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LOW NOISE DEVICES FOR SATELLITE COMMUNICATION SYSTEMS

by

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SUMMARY

This Report outlines the current state of the art in RF low noise front end components for use in satellite communication receivers. *To illustrate the* importance of the receiver front end the various noise components in the complete communication system are described. Some typical examples of low noise receivers are given.

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LIST OF CONTENTS

1 INTRODUCTION

In recent years very rapid developments in silicon and gallium arsenide materials and device technology have led to considerable improvements in the parameters of microwave receiver front end components. In particular the improvement in gallium arsenide technology has led to the availability of field-effect transistors (FET) whose low noise performance is approaching that achievable with the parametric amplifier (paramp). Similarly, at higher frequencies, improvements in GaAs Schottky barrier diodes has led to the development of low loss mixers which, when used with low noise IF amplifiers, can result in a low noise front end which may also rival the paramp.

Thus the early predominance of the paramp in the lowest noise systems is now being challenged by the transistor amplifier at frequencies below 20-25 GHz and by the mixer at frequencies above 25 GHz. The low noise receiver designer concerned with satellite borne, mobile, or ground station receivers is therefore increasingly faced with a choice between competing components. It is the purpose of this present Report to summarise the current state of the art in RF front end component development. The system/receiver designer may therefore be helped to put into perspective the confusion of choice between RF front ends which appears to exist at some frequencies.

The satellite communication system will be briefly examined in order to illustrate the relative importance in the system of the receiver front end. The frequency range of approximately 1-40 GHz only will be considered. The components of the established superheterodyne receiver will be subsequently described and the state of the art development in each component will be reviewed. Some typical examples of low noise front ends will be discussed in order to illustrate the component choice facing the receiver designer.

2 THE SATELLITE COMMUNICATION SYSTEM

The satellite communication system operates through the use of the satellite as a relay or repeater. Information transmitted up to the satellite from a ground station or mobile transmitter may be amplified and retransmitted to the ground at the same or at a different frequency. Global coverage may be achieved in this retransmission by a geostationary satellite. Alternatively, synchronous satellites may achieve regular coverage of a chosen geographical area. The basic requirement of the up-down link is that the system should enable information of acceptable quality to be relayed. 'Acceptable quality' of course depends upon the type of information to be relayed. The quality of the information is usually defined

in terms of a signal-to-noise ratio. The characteristics of most RF modulation methods suitable for use in multi-channel satellite communication links are such that the baseband signal-to-noise ratio is directly proportional to the RF carrier-to-noise ratio provided that the carrier-to-noise ratio is sufficiently larger than unity.

Thus for the purposes of this Report, that is an examination of the importance and parameters of the receiver front end, only information quality as expressed by the carrier-to-noise power ratio (C/N) will be considered. First however the simple characterisation of noise will be introduced.

2.1 Noise characterisation

Consider a resistance R at an absolute temperature T. Owing to the random motion of electrons a thermal noise voltage is generated across its open circuit terminals. The available noise power from the resistor is given by[1]

$$P_{av} = kTB \tag{2-1}$$

where k = Boltzmann's constant and B is the bandwidth over which the noise voltage is measured. Equation (2-1) leads to the procedure of referring to a *noise power in terms of a noise temperature* even though the noise power may have its origin in a physical mechanism other than the thermal agitation of electrons in a hot resistance. The noise 'temperature' at a port may be defined as the temperature of a passive system having an available noise power per unit bandwidth equal to that of the actual port at a specified frequency.

It follows that the total output noise power of a system can be represented in terms of an operating noise temperature T_{op}. The operating noise temperature is given by the sum of the noise temperatures of the noise powers arising within the system. Thus the total output noise power of a single response receiver is expressed by

$$\begin{aligned} N_{TO} &= GkBT_{op} \\ &= GkB(T_i + T_e) \end{aligned} \tag{2-2}$$

where G is the receiver gain, T_i the noise temperature of the input termination, and T_e the (average) "effective input noise temperature" of the receiver. T_e is defined as the temperature of a noise source at the input of a noiseless receiver that would result in the same noise power at the receiver output as that of the actual receiver when connected to a noise free input termination. 076

2.2 System C/N

The communication system comprises both up and down links and both links contribute to the final signal-to-noise ratio seen by the receiver demodulator. The ground transmitting station initially transmits with a finite carrier-to-noise ratio. Noise generated within the satellite uplink receiver combines with noise and intermodulation products generated within the satellite transmitter to produce a finite downlink carrier-to-noise ratio. This carrier-to-noise ratio is maintained down to the receiver station antenna input. Additional noise enters the system via the receive antenna and the receiver contributes its own inherent noise to the carrier.

The received carrier power at the input to satellite or earth station receiver may be written

$$C(\mathrm{dBW}) = (P_{gt} - L_{gf} + G_{ga}) - (L_{fs} + L_a) + G_{ra} \qquad (2\text{-}3)$$

where P_{gt} = transmitter output power in dBW

L_{gf} = transmitting antenna feed system loss in dB

G_{ga} = transmitting antenna gain in dB

L_{fs} = free-space loss in dB

L_a = attenuation by atmosphere in dB

G_{ra} = receiving antenna gain in dB.

The gain of an antenna is given by

$$G_a = \frac{4\pi A_{eff}}{\lambda^2} \qquad (2\text{-}4)$$

where A_{eff} is the effective area, or aperture, of the antenna, and λ is the carrier wavelength.

Free-space loss is given by

$$L_{fs} = \left(\frac{4\pi r}{\lambda}\right)^2 \qquad (2\text{-}5)$$

where r is path length between the transmitter and receiver.

The expression in the first set of brackets on the right hand side of equation (2-3) is called the 'effective isotropically radiated power' (EIRP) of

the transmitting station. The system noise contributions may be taken into account by assigning each noise power a noise temperature. Noise in the uplink may be characterised by temperature $T_{u\ell}$. Noise and intermodulation noise added by the satellite transponder may be characterised by a temperature T_{st} and the downlink noise by temperature $T_{d\ell}$.

The carrier-to-noise ratio at the earth station receiver input is

$$C/N_i(\text{dB}) = \text{EIRP}_{sat} - (L_{fs} + L_a) + G_{ra} - 10 \log\left[k(T_{u\ell} + T_{st} + T_{d\ell})B\right] \quad . \qquad \text{... (2-6)}$$

The downlink noise temperature $T_{d\ell}$ includes the receiver effective input noise temperature. The bracketed term $T_{u\ell} + T_{st} + T_{d\ell}$ is thus the ground station receiver operating noise temperature T_{op} of equation (2-2). Including the receiver noise temperature in equation (2-6) essentially means that the receiver can be considered noiseless so that equation (2-6) also gives the C/N at the output of the ground station receiver.

The downlink noise temperature $T_{d\ell}$ is usually the dominating component of T_{op}. Rewriting equation (2-6) yields

$$C/N(\text{dB}) = \text{EIRP}_{sat} - (L_{fs} + L_a) + G_{ra} - L_{rf} - 10 \log\left[k(1 + r)T_{d\ell}B\right] \qquad \text{... (2-7)}$$

where $r = (T_{u\ell} + T_{st})/T_{d\ell}$.

The system carrier-to-noise ratio is thus determined, for a given satellite EIRP by G_{ra} the receiving antenna gain, and $T_{d\ell}$ the downlink noise temperature.

2.3 The system noise components

The system noise components have been grouped into uplink, satellite transponder, and downlink components, characterised by $T_{u\ell}$, T_{st} and $T_{d\ell}$ respectively. Since $T_{d\ell}$ is normally the dominating component $T_{u\ell}$ and T_{st} will only be briefly considered.

2.3.1 Uplink noise

This noise arises basically from both phase and amplitude noise present on the original carrier transmitted up from the ground station together with noise entering the earth pointing satellite antenna from the earth and its atmosphere which acts as a noise source. Obviously a higher EIRP from the ground station transmitter is required to minimise the effect of the extraneous noise from the 076

earth and atmosphere. Carrier transmitted noise is however a function of the ground transmitter design.

2.3.2 Satellite transponder noise

An expression similar to equation (2-7) determines the C/N at the input to a (noiseless) satellite transponder. The satellite receiver noise temperature, intermodulation noise (assumed similar to a white noise spectrum for a large number of channels in a frequency multiplexed system), frequency translation effects and transmitter noise combine to produce noise from the satellite.

The satellite receiver noise temperature must of course be minimised although the point of diminishing returns is met when the receiver noise is small compared with the remaining noise components. Receiver noise reduction is dealt with in the following section. Intermodulation effects and transmitter noise are fundamental properties of the transmitter design and the type of output power amplifier that is used. Some degree of power amplifier back-off is normally utilised to minimise intermodulation noise.

2.3.3 Downlink noise

Downlink noise, as characterised by $T_{d\ell}$ in equation (2-7), comprises two important components; noise entering the ground station receiver via the antenna from extraneous sources other than the satellite, and noise added by the receiver. The two components are discussed separately.

(a) Receiving antenna noise temperature

This noise input comprises contributions from sky noise and noise arising from losses within the antenna. The antenna noise temperature T_a is usually defined as the noise temperature measured at the feed input, *ie*

$$T_a = T_s + T_m \tag{2-8}$$

where T_s is the sky and atmosphere noise contribution and T_m the contribution due to antenna losses. T_s is shown plotted in Fig 1 as a function of frequency for different antenna elevation angles[2]. The horizontal elevation of course represents the maximim depth of atmosphere seen by the antenna. The two peaks at 22 GHz and 60 GHz are due to water vapour and oxygen absorption respectively. Re-emission of absorbed energy constitutes noise.

Fig 1 in fact shows the contribution of sky noise to the antenna temperature for an ideal antenna, *ie* finite main lobe and no side lobes. Noise may however

enter the system via a side lobe. The noise temperature added in this way is reduced by the ratio of the main to side lobe gain. The noise temperature of the antenna could however be dominated by a side lobe contribution should the side lobe point towards a particularly strong noise source, *eg* the sun. The temperature of the sun[3] is shown in Fig 2. T_m arises from ohmic losses within the antenna reflector, supports, and primary feed. This lost power can reradiate noise back into the antenna. The efficiency of the antenna and therefore this contribution to noise is, unlike T_s, a function of the antenna design. A typical value of T_m is 12 K at 3-4 GHz for a 30 metre Cassegrain antenna at 5^o elevation.

(b) Receiver noise temperature

Fig 3 shows a conventional receiver comprising a basic input stage of RF amplification, the RF front end, followed by a heterodyne frequency down-converter and an IF amplifier.

The essential role of the receiver is to down-convert the RF carrier incident on the antenna to an IF which may then be demodulated and the baseband signal retrieved. The receiver must perform the down-conversion while contributing the minimum possible noise to the system. The overall receiver noise level is represented by the effective input noise temperature T_e of equation (2-2). T_e may be calculated provided the effective input noise temperatures and available gains of the receiver components are known. The noise performance of the heterodyne receiver is however complicated by the fact that the receiver can have more than one response. Provided that the RF amplifier bandwidth is sufficiently large, then RF frequencies incident on the antenna which are spaced by the magnitude of the IF frequency on either side of the local oscillator frequency, will be translated to the IF frequency. The situation is shown in Fig 4.

A double response receiver is designed for $G_u = G_\ell$ and a single response receiver is designed for either $G_\ell >> G_u$ or $G_u >> G_\ell$. The receiver operating noise temperature given by equation (2-2) is for single response operation. If the receiver has more than one response then the total receiver output noise power is given by[1]

$$N_{TO} = G_1kB_1(T_{i_1} + T_e) + G_2kB_2(T_{i_2} + T_e) + \ldots G_nkB_n(T_{in} + T_e) \qquad (2\text{-}9)$$

where G_n, B_n, and T_{in} refer to the gain, bandwidth and input termination temperature of the nth response.

076

For the double response receiver, assuming that the input termination temperature is the same for both responses and that the bandwidth is determined by the IF amplifier, the receiver output noise power is given by

$$N_{TO} = kB_{IF}(T_i + T_e)\left[G_\ell + G_u\right] \tag{2-10}$$

and if $G_\ell = G_u$

$$N_{TO} = 2GkB_{IF}(T_i + T_e) \tag{2-11}$$

or the operating noise temperature is twice that of the single response receiver given by equation (2-2). Obviously the use of a double response receiver for single response reception results in a receiver operating noise temperature of twice that obtainable with a single response receiver and therefore a reduction of 3 dB in receiver sensitivity.

Most satellite communication systems make use of single response receivers, consequently signals and noise present in the unwanted sideband, or 'image' response, must be suppressed. This is usually achieved by incorporating a selective filter placed in the receiver cascade just prior to the mixer. Alternatively, or additionally, a mixer with inherent image suppression properties is used. Some applications however, *eg* radioastronomony, may require double response receivers where broadbandwidth is required and the signal is essentially broadband noise.

Considering then the single response receiver, the operating noise temperature is given by equation (2-2). The effective input noise temperature of the receiver may now be calculated. The RF amplifier is characterised by an available gain G_r and an effective input noise temperature T_{er}. Similarly the IF amplifier is characterised by G_i and T_{ei} and the mixer by G_m, its conversion gain, and T_{em}. The demodulator and following post receiver may also be characterised by an effective input noise temperature T_{ed}. The well known noise cascade equation[1] thus yields, for the cascade of matched components shown in Fig 3, the effective input noise temperature of the receiver

$$T_e = T_{er} + \frac{T_{em}}{G_r} + \frac{T_{ei}}{G_r G_m} + \frac{T_{ed}}{G_r G_m G_i} . \tag{2-12}$$

It is evident from equation (2-12) that the receiver noise temperature is
076 minimised when the input (front end) amplifier has simultaneously a high gain and a low effective noise temperature.

(c) Effects of loss between the antenna and the receiver

It may not be possible in some applications to connect the receiver directly to the antenna feed. Quite often it is necessary to protect the input to the RF amplifier from strong signals which may fall within the amplifier passband and cause amplifier damage or spurious outputs within the IF bandwidth. A filter or limiter may therefore be introduced between the antenna and the amplifier. Similarly the antenna may be connected to the RF amplifier via a diplexer or duplexer if the receiver forms part of a transponder, or by a coupler if receiver noise or signal tests are required for field monitoring of the receiver performance. In some circumstances a limiter, diplexer and coupler may all be present. All three components will introduce some degree of loss between the antenna and the receiver. Loss may also be introduced if the receiver front end is connected via a long length of waveguide or coaxial cable to the antenna. For illustrative purposes all the losses identified thus far may be represented by an attenuator L_ℓ at an ambient temperature T_0 connected between the antenna and the receiver.

The effective input noise temperature of an attenuator is given by[1]

$$T_{e\ell} = (L_\ell - 1)T_0 \ . \qquad (2\text{-}13)$$

Thus application of the noise cascade equation with an attenuator L_ℓ between the antenna and receiver of Fig 3 yields

$$T_e = T_{e\ell} + T_{er}L_e + \frac{T_{em}L_e}{G_f} + \frac{T_{ei}L_\ell}{G_rG_m} + \frac{T_{ed}L_\ell}{G_rG_mG_i} \qquad (2\text{-}14)$$

where $T_{e\ell}$ is given by equation (2-13) and T_e the effective input noise temperature of the receiver is measured at the input to the attenuator L_ℓ.

It is obvious from equation (2-14) that any loss mechanisms that occur between the antenna and receiver must be minimised. The receiver effective input noise temperature T_e given by equation (2-12) is increased by the factor L_ℓ. The attenuator ambient temperature T_0 is also a noteworthy parameter.

(d) Effects of atmospheric loss

Operating experience gained with geostationary satellites has shown that carrier fading and an increase in system noise has occurred primarily during periods of rain. As might be expected this absorption is frequency dependent and is also dependent on rainfall rate which is related to drop size distribution. 07

It is in practice difficult to predict accurately the effective loss due to rain over a slant path through the atmosphere as there is invariably a lack of detailed meteorological information in both horizontal and vertical directions.

The effects of rain absorption and the consequent increase in system noise can be included in the receiver operating temperature. Let L_r = rain absorption factor and T_{rain} = the effective temperature of the rain. The rain can be considered as an attenuator positioned between the antenna and the source of sky noise. Consequently the receiving antenna noise temperature can be written

$$T_{ar} = T_m + \frac{T_s}{L_r} + \frac{(L_r - 1)}{L_r} T_{rain} \tag{2-15}$$

where T_{ar} is again measured at the feed input. Fig 5 shows the attenuation due to absorption by rain as a function of surface rainfall rate[4] at 4 and 6 GHz assuming an 18 km long equivalent rain bearing distance. Fig 6 shows the consequent relative noise temperature increase, for a nominal receiver noise temperature in the absence of rain of 50 K, at 4 GHz when the rain noise temperature is assumed to be 290 K[4]. The situation may be expected to deteriorate at higher signal frequencies, *ie* 11 and 14 GHz[5].

2.4 Receiver G/T

Equation (2-7) shows that for a given satellite EIRP the system carrier-to-noise is determined by the receiving antenna gain, given by equation (2-4), and by $T_{d\ell}$, the downlink noise temperature.

$T_{d\ell}$ is essentially the operating noise temperature of the ground station receiver if the uplink and transponder noise are small compared to $T_{d\ell}$, *ie* if $r \ll 1$ in equation (2-7). Thus $T_{d\ell} = T_{op}$ of equation (2-2). The receiver input termination temperature $T_i = T_a$ of equation (2-8) in the absence of rain or $T_i = T_{ar}$ of equation (2-15) in the presence of atmospheric absorption due to rain. The receiver operating noise temperature is thus

$$T_{op} = T_{ar} + T_e \tag{2-16}$$

where T_e, the effective input noise temperature of the receiver is given by equation (2-14). It is important to note that the operating noise temperature expressed by equation (2-16) is that as measured at the feed input of the antenna. The receiver operating noise temperature could alternatively be specified at the
76 input to the receiver front end amplifier. T_{ar}, the antenna noise temperature given by equation (2-15), would then have to be modified to take into account the

attenuation of any additional component or loss that occurs between the antenna and the input to the receiver. The carrier power would similarly have to be specified at the receiver front end. The ratio between the receiving antenna gain and the receiver operating noise temperature G_a/T_{op} is used as a ground station, or any receiving station, figure of merit and obviously, from equation (2-7), maximising this ratio for a given transmitter EIRP maximises the C/N ratio at the receiver demodulator input. The ground or receiver station designer's task is thus to maximise the ratio G_a/T_{op}, or for a given carrier frequency which determines the level of sky noise, to minimise T_e, the receiver effective input noise temperature and to maximise the receiving antenna gain.

The first and most obvious trade-off is between G_a and T_e and this is ultimately a matter of cost. The cost of an antenna increases with increasing diameter, *ie* with increasing antenna gain, while generally the cost of low noise receivers increases with decreasing noise temperature. Fig 7 shows receive antenna cost, including mount, feed and drive, as a function of antenna diameter while Fig 8 shows the cost of front end RF amplifier units as a function of system noise temperature requirements. Both graphs are constructed from relatively sparse data[6], corrected for current costs, and are essentially illustrative only.

It is not however always possible to select either a high gain antenna or a low noise receiver as a simple cost trade-off exercise. Quite often the receiving station antenna gain must be chosen to enable easy satellite acquisition. For example, a mobile receive station such as a ship or aircraft would normally require a low gain antenna, *ie* a wide antenna beamwidth to ensure that the satellite transmissions are not lost during the receive station platform motion. Domestic TV installations receiving from satellites may be constrained to have small diameter antennas for environmental reasons or to avoid the necessity for moving the antenna to accommodate small excursions of the satellite from the nominal geostationary position. In such applications the receiver G/T must be maximised by the judicious choice of receiver components in order to minimise T_e. However the point at which T_e becomes equal to the antenna noise temperature T_a represents the onset of 'diminishing returns' in cost terms as the receiver operating temperature T_{op} will be determined by natural effects outside the control of the receiver design.

The trade-off between G_a and T_e will be dealt with in a later section. First the receiver components, in particular the RF front end, will be discussed in order to illustrate the effect on T_e of the choice of the state of the art 076
components in the receiver cascade.

3 RECEIVER FRONT END COMPONENTS

Equation (2-14) shows that any loss mechanism that lies between the antenna and the RF front end component of the receiver will contribute to the receiver noise temperature. Attenuation due to long lengths of waveguide or coax can of course be minimised by mounting the RF amplifier as close as possible to the antenna. Losses due to protection devices and filters etc will be dealt with in the last section or as they arise in the discussion of the various front end components.

Although front end components, *ie* the RF amplifiers, will be discussed, the receiver noise temperature is, as illustrated by equation (2-14), not determined by the RF amplifier alone. The 'second stage' or downconverter contribution can be significant if the front end amplifier gain is low or the downconverter noise level is high. Fig 9 shows how the downconverter noise temperature contribution to T_e depends upon the RF amplifier gain for different values of the downconverter noise temperature. The graphs are calculated using equation (2-12), but omitting the last term.

Obviously for the lowest noise systems the downconverter noise temperature must be minimised. However the most important component from the noise viewpoint in the receiver cascade is the RF amplifier which must itself add the minimum amount of noise and also have sufficient gain to minimise the downconverter contribution.

However apart from the noise and gain parameters of the RF amplifier the dynamic range of the amplifier is an important consideration. Dynamic range is limited at the low signal end by the noise performance of the amplifier and at the large signal end by the onset of nonlinearity due to amplifier gain compression. The upper limit of the dynamic range is characterised either by the input or output level, the saturation power output, at which 1 dB gain compression occurs, or by the intermodulation intercept point[7]. The intermodulation intercept point is measured by applying two signals f_1 and f_2 in the passband of the amplifier. If the amplifier is not linear the signals will mix producing 2nd order intermodulation products at $2f_1$, $2f_2$, $f_1 + f_2$ and $f_1 - f_2$. With amplifier bandwidths of less than an octave these effects will probably be negligible. 3rd order products will occur at $2f_1 - f_2$, $2f_1 + f_2$, $2f_2 - f_1$ and $2f_2 + f_1$ where $2f_1 - f_2$ and $2f_2 - f_1$ will fall within the amplifier passband.
76 The output power of f_1 and f_2 and the 2nd and 3rd order intermodulation products are plotted as a function of the equal input powers f_1 and f_2. The

intercept point occurs at the output power at which the levels of f_1 and f_2 and the intermodulation product powers (usually 3rd order products) are equal. The intercept point is obtained in this manner by extrapolation of the linear part of the input/output power plot for f_1 and f_2.

The following sections outline the various RF front end options open to the receiver designer.

3.1 Travelling-wave-tube amplifiers (TWTA)

Travelling-wave-tube amplifiers can be designed to operate as low noise amplifiers (LNTWTA). *TWTs essentially consist of a periodic interaction structure* and an electron beam where the phase velocity of the signal on the interaction structure is approximately equal to the electron velocity in the beam. Energy can thus be continuously transferred from the electron beam to the RF travelling wave and amplification takes place. TWTs use a helix as a slow wave interaction structure and since the electron beams are quite long permanent magnets are used to focus the beam. Fig 10 shows a schematic of a TWT. LNTWTs may be constructed by optimising the design of the input coupling, gun construction and focussing.

Noise performance can be traded off against bandwidth, weight, etc. *Table 1 shows a selection of the currently available* LNTWTAs.

Table 1

Currently available LNTWTAs

Frequency range GHz	Small signal gain dB	Noise temperature K	Weight lb
1.0-2.0	25	615	17
1.2-2.4	25	440	18.7
2.2-2.3	25	400	17
2.7-3.2	25	440	15
2.0-4.0	25	750	17
2.0-8.0	25	1350	18
3.4-3.6	25	515	19.5
3.7-4.2	25	615	17
4.0-12.0	25	1800	18
4.4-5.25	25	615	11.0
4.8-9.6	40	2300	5.5
5.85-7.25	30	1000	7.5
7.0-11.0	25	1000	8.5
8.0-12.0	25	1180	18
8.0-16.0	30	2300	18
12.0-18.0	30	1800	6.5
18.0-26.0	25	5500	18
26.0-40.0	25	7000	25

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TWTs are rugged and can withstand input RF spikes of up to 1 kW without damage and therefore do not require input protection when used in communication receivers. They are capable of wide dynamic range (IM intercept point +10 to +20 dBm) operation and the very wide bandwidths available with these devices make them suitable for use in active receivers, *eg* in ECM applications. They can be supplied with integral preselect bandpass filters and an integrated power supply. Power supply requirements are about 20 W. Typical size of a LNTWTA is approximately 3 in × 3 in × 13 in. Tube life, limited by cathode wearout, is estimated at 10^4-10^5 hours. The particular advantages of the LNTWTA are wide bandwidth, well defined package size, and the fact that they are readily available. New developments are however at a very low level with high frequency ECM systems as perhaps the prime user. Where possible the LNTWTA is being replaced by the more compact and more reliable GaAs FET amplifier. LNTWTAs remain attractive at the higher frequencies >20 GHz for octave bandwidth applications and where GaAs FET amplifiers cannot yet compete.

3.2 The maser

The maser amplifier is perhaps of limited importance to the modern satellite communication receiver designer. Nevertheless it is mentioned because it is capable of the lowest noise performance of all of the front end amplifiers discussed in this section and could find application in deep space telemetry links.

Basically the maser relies on the stimulated emission of radiation from a crystal, *eg* ruby. Three electronic energy levels are required and impurities such as chromium are used to provide the intermediate energy level in ruby. A source of RF power, the pump, is used to cause electron population inversion by absorption, from the lowest to the highest level. Subsequent transition from the highest energy level to the intermediate level results in amplification of an RF signal of frequency corresponding to the energy difference between the highest and intermediate energy level. The lower the crystal ambient temperature, the greater the degree of population inversion that results from a given pump power and the smaller will be thermal disordering effects and spontaneous emission noise. Masers usually operate at ambient temperatures of about 4 K. The energy level differences, and consequently the signal frequency, can be adjusted by a magnetic field.

A cavity maser is constructed by placing the crystal in a cavity which is
076 resonant to both signal and pump frequencies. The required low operating temperature is obtained by immersing the cavity in liquid helium. For a sufficiently

high pump power the cavity input admittance has a negative real part at the signal frequency. The maser is thus used as a circulator coupled negative resistance amplifier.

A travelling wave maser is constructed by mounting the crystal in a transmission line slow wave structure which supports both signal and pump frequencies. The slow wave structure is placed in a uniform magnetic field which can be adjusted to give operation at the required frequency and cooled to about 4 K. The gain of the TWM is directly proportional to its physical length. The bandwidth is determined fundamentally by the paramagnetic resonance linewidth, *ie* the frequency range over which radiation interacts with the active material which fact tends to make the maser a rather narrow band amplifier. Bandwidth can be increased at the expense of gain by making the dc magnetic field slightly nonuniform or by stagger tuning, *ie* by cascading amplifier sections with slightly different magnetic field strengths and hence signal frequencies. Each section would however require a different pump frequency if the bandwidth is to be increased beyond about 50 MHz.

The noise temperature of the maser is given by the product of the inverse of the population inversion ratio, typically 2-3, and the maser material ambient temperature. For a ruby crystal at 4 GHz maser action is obtained with a net gain of 33 dB and a bandwidth of 8 MHz for a magnetic field of 25×10^4 A-t/m and a pump frequency of 30 GHz. The maser noise temperature was 4.5 K for a material ambient of 4.2 K. In practice the losses of the input feed line would dominate the amplifier noise temperature. Fig 11 shows the noise temperatures of masers reported in the literature and summarised in Ref 8.

Signal frequencies between 1 and 40 GHz require pump frequencies between 20 and 80 GHz. Pump powers required vary between 50 mW for a 4 GHz maser and 500 mW for a broadband or high frequency maser. Masers are extremely stable amplifiers and are relatively insensitive to variations in pump power and frequency. Nevertheless the procurement of pumps of sufficient power at the highest frequencies could be a problem. X-band klystrons and high efficiency triplers are perhaps the only solution at the present time.

Early masers required the use of large permanent magnets. Later developments however make use of the maser liquid helium cold bath to operate superconducting electromagnets so that some saving in weight is achieved. The particular disadvantages of the maser are cost, compexity, relatively narrow bandwidth, <5%, limited dynamic range (IM intercept point < -30 dBm), and the requirement 076
for cryogenic refrigeration.

3.3 Tunnel diode amplifier (TDA)

Tunnel diode amplifiers are negative resistance amplifiers operated in a circulator coupled reflection mode. The negative resistance arises from the negative slope region of the tunnel diode dc current-voltage characteristic. The device is a degenerately doped pn junction which displays a tunnelling limited current at low bias levels and an exponentially increasing thermal diffusion current at higher bias levels. The tunnelling current initially rises with increasing bias, peaks and then begins to fall giving rise to the negative slope. The current through the device is dominated by thermal diffusion current as the bias increases and the current voltage characteristic resumes its positive slope.

Noise in the TDA arises from three contributions, two of which arise from the tunnel diode itself. The first arises from resistive losses in the circulator and in the stabilisation and tuning circuitry surrounding the device. The diode noise contributions arise from Johnson or thermal noise generated in the diode series resistances and shot noise generated by the dc bias current flowing through the diode. The principal noise contribution is shot noise although at high frequencies, *ie* 25 GHz, the thermal noise contribution dominates.

The three most important semiconductor materials used for the fabrication of tunnel diodes are GaSb, Ge and GaAs. The major materials dependent electrical differences between the diodes are noise and RF dynamic range. The range of diode forward bias over which the negative slope exists of course determines the amplifier RF dynamic range. For GaSb, Ge and GaAs diodes the range is respectively 225 mV, 290 mV and 440 mV. GaAs diodes thus have the widest dynamic range. The tunnel diode noise temperature is given approximately by K.300 K where K is the shot noise constant of the semiconductor material. For GaSb, Ge and GaAs K is respectively 0.85, 1.2 and 2.4. GaSb diodes thus have the lowest noise temperature but unfortunately have the smallest dynamic range. Ge diodes are usually used as the best compromise between dynamic range and noise.

Practical TDAs have noise temperatures between 380 K at 1 GHz and 800-1000 K at 20 GHz with an IM intercept point of between -30 to -10 dBm. Operation above 25 GHz is dominated by the diode series resistance. The diode noise temperature increases rapidly with increasing frequency as the thermal noise generated in the series resistance dominates and the diode negative resistance is reduced as the frequency increases. The highest quality tunnel diode has a resistive cut-off frequency in excess of 50 GHz (*ie* the frequency above which the diode no longer exhibits a negative resistance). This frequency thus represents the absolute

limit of operation of the TDA, however the use of the TDA as a practical low noise amplifier is usually limited to approximately 25 GHz or below.

The TDA is basically a simple amplifier to construct and is readily produced in microstrip form. It thus has the advantage of small size and in addition has the advantage of low power drain. TDAs have therefore found application in some satellite transponder receivers, *eg* in Intelsat IV. Broad bandwidths are possible with up to a full octave practical. The major disadvantage of the TDA is the low RF dynamic range. It is steadily being replaced by the GaAs FET amplifier at frequencies up to 14 GHz and by miniature parametric amplifiers and mixers at higher frequencies. Since the important electrical parameters of the TDA depend fundamentally on the constants of the diode semiconductor material, little further improvement of either noise or RF dynamic range is currently expected. Table 2 shows a selection of the currently available TDAs.

Table 2

Currently available LNTDAs

Centre frequency GHz	Noise temperature K	Bandwidth MHz
1.6	530	100
2.2	530	100
3.95	530	500
7.5	530	500
12	630	1000
15	860	500
18	860	2500
30	1550	2500

The TDA should however not be completely discounted as a possible receiver front end amplifier in a future 20-30 GHz communication link. The practical difficulty will be in developing a tunnel diode with a resistive cut-off frequency of approximately 90 GHz. The simplicity and low power drain of the TDA may provide sufficient impetus for such a development in a satellite borne application.

3.4 Parametric amplifiers (paramps)

Parametric amplifiers depend for their operation on the nonlinear capacitance-voltage characteristic of semiconductor pn junction diodes. The diodes, designed for the purpose are called varactors. The varactor capacitance varies greatly with bias (~100% variation) and the diode is continually 'pumped' through its characteristic by a high level, high frequency oscillator. The varactor is

embedded in a microwave structure which supports the pump frequency, the signal frequency and the 'idler' frequency, normally the difference between the pump and signal frequencies, *ie* $(f_p - f_s)$. Power gain is achieved by extracting energy from the pump source at either the signal frequency, when the device is called a parametric amplifier, or at the idler frequency when both amplification and frequency conversion occur and the device is called a parametric converter. A schematic of a three frequency parametric device is shown in Fig 12.

The operation of parametric devices is described by the Manley-Rowe equations[9], a general set of equations relating power flowing into and out of an ideal nonlinear reactance. When the device is to be operated as a parametric amplifier the idler frequency f_i is chosen to be the difference between the pump and signal frequencies $(f_p - f_s)$. Application of the Manley-Rowe equations[10] shows that the relationship between the signal, pump and idler powers and frequencies is given by

$$\frac{P_s}{f_s} + \frac{P_p}{f_p} = 0 \tag{3-1}$$

and

$$\frac{P_i}{f_i} + \frac{P_p}{f_p} = 0 \ . \tag{3-2}$$

If pump power is supplied to the varactor then P_p is positive. Thus both P_s and P_i are negative, *ie* power is supplied from the varactor to the signal generator at frequency f_s and to the idler circuit at frequency f_i. If gain is defined as the ratio of power supplied to the generator resistance at f_s by the varactor to that supplied by the generator to the varactor then it is evident that infinite gain is possible since power can be supplied to the generator irrespective of whether the generator is supplying power to the varactor. The pumped varactor thus has a negative resistance as part of its signal impedance. A detailed analysis of the parametric amplifier[10] shows that the resistance component of the signal circuit impedance is given by

$$R_{sig} = -\frac{\gamma^2}{4\pi^2 f_s f_i C_0^2 R_i} \tag{3-3}$$

where f_s and f_i are the signal and idler frequencies, C_0 is the varactor
076 capacitance at the bias point, R_i is the idler circuit loss resistance and γ, given below, is the varactor modulation parameter, *ie*

$$\gamma = \frac{1}{2}\left[\frac{C_{max} - C_{min}}{C_{max} + C_{min}}\right] . \qquad (3\text{-}4)$$

It is evident from equation (3-3) that if the idler circuit was not present, *ie* $R_i = \infty$ a negative resistance would not have resulted. An alternative form of equation (3-3) results from the inclusion of the varactor diode cut-off frequency $\omega_c = \frac{1}{C_0 R}$ where R is the varactor series resistance. Equation (3-3) becomes

$$R_{sig} = -\frac{\gamma^2 f_c^2 R^2}{f_s f_i R_i} \qquad (3\text{-}5)$$

which equation illustrates the figure of merit for the varactor diode, *ie* the product γf_c which should be maximised.

The negative resistance paramp is usually operated in the circulator coupled reflection mode. If the device in Fig 12 is to be operated as an upconverter the idler or output frequency is chosen to be the sum of the pump and signal frequency, *ie* $(f_p + f_s)$. The Manley-Rowe equations yield

$$\frac{P_s}{f_s} + \frac{P_0}{f_0} = 0 \qquad (3\text{-}6)$$

$$\frac{P_p}{f_p} + \frac{P_0}{f_0} = 0 . \qquad (3\text{-}7)$$

The gain of the upconverter is thus P_0/P_s which from equation (3-6) is simply f_0/f_s. Consequently to achieve a reasonable gain f_p must be considerably larger than f_s. As a result the output frequency will be correspondingly higher than the input frequency. The advantage of the parametric upconverter is that its input resistance is positive and it can therefore be operated without a circulator. However the device is not often used and becomes unattractive at signal frequencies in excess of 1 GHz. In order to achieve a worthwhile level of gain the output frequency will be 10× or greater than the input frequency and this requires a pump source of frequency many times the signal frequency. Subsequent down-conversion becomes increasingly lossy as frequency increases so that the second stage noise contribution in a receiver with a parametric converter front end will 076
be significant unless the converter gain is sufficiently high. This is only

achieved at the expense of a higher output frequency, *ie* a higher pump frequency which increases the problem.

The parametric devices briefly discussed so far are in theory capable of low noise operation since the ideal nonlinear reactance does not contribute thermal noise to the circuit. In practice however thermal noise will be added by the inevitable series resistance associated with the practical varactor diode which arises from contact resistance and spreading resistance within the semiconductor material. Other forms of noise are not significant in the reverse biased varactor diode provided they are not pumped to the extent that they are driven into forward conduction or reverse breakdown. The noise temperature of the negative resistance paramp is given by[10]

$$T_e = \left(1 - \frac{1}{G}\right) T_A \frac{\left[1 + \frac{f_s}{f_i}(\gamma f_c)^2\right]}{\left[\frac{(\gamma f_c)^2}{f_s f_i} - 1\right]} \tag{3-8}$$

where T_A is the paramp ambient temperature and G is the paramp power gain, given by

$$G = \frac{4R^2_{sig}}{R_G - R_{sig}} \tag{3-9}$$

where R_G is the sum of the input line impedance and the varactor loss resistance. Equation (3-9) is calculated assuming a high gain approximation. It is evident from equation (3-8) that the paramp noise temperature may be reduced by either operating at a low ambient temperature T_A or by operating with a high idler frequency, *ie* 'electronic cooling' the paramp. An optimum idler frequency can be calculated from equation (3-8). T_e is a minimum when $f_i = \gamma f_c$, *ie*

$$T_{e_{min}} = T_A \frac{2f_s}{\gamma f_c} \; . \tag{3-10}$$

This minimum noise temperature is not practically attainable when the highest quality varactor diodes are used due to the problem of providing a pump at a frequency comparable with γf_c. Thus the paramp noise temperature is determined
76 by the idler frequency and the paramp ambient temperature.

The pump frequency may well be chosen with the minimisation of the noise temperature in view. However pump power must also be taken into account as power is often the limiting feature of pumps. The minimum voltage drive to the varactor is of the order of 1 V peak to peak in order to achieve a suitable C_{max} to C_{min} swing. Providing this voltage from a 50 Ω impedance implies that the minimum pump power is 2.5 mW. However this value does not take into account diode losses or pump matching circuit losses. Practical pump powers are in the range 10-60 mW. The pump output power must be controlled accurately as it is evident from equations (3-3) and (3-9) that fluctuations in γ lead directly to gain variations. The size of the negative resistance is, through γ, directly dependent on the pump power level. The sensitivity of the paramp gain to fluctuations in the pump power level increases rapidly with increasing gain. Consequently paramp stages are rarely operated at gains in excess of 15 dB for wideband (>10% fractional bandwidth) applications or 20 dB for narrowband (<5% fractional bandwidth) applications. At a gain of 20 dB a 0.1 dB change in pump power will usually produce a paramp gain change of at least 1 dB. The paramp is quite different from the maser in this respect. The state of the art in solid state pumps is shown in Fig 13 where the output power of fundamental GaAs Gunn and Si Impatt oscillators is shown as a function of frequency.

Paramp gain saturation is, to a first order, limited by the available pump power. As the input signal power increases the pump must convert more and more power to the signal and idler frequencies until it eventually limits. Practically, good paramps have 1 dB gain compression points at a signal input power of -20 dBm with -30 to -35 dBm being typical. Dynamic range is unfortunately rather limited when compared with the LNTWTA and transistor amplifiers. One of the disadvantages of the paramp which utilises single tuned circuits, *ie* in the signal and idler circuits, is narrow bandwidth. The amplifier may be regarded as an impedance terminating a transmission line. The bandwidth is limited by the reactance, tuned out at the signal frequency, seen at the input which increases towards the required band edges. Broadbanding the input line is carried out by adding appropriate reactances in the signal line at the correct distance from the diode. The single tuned paramp follows the gain-bandwidth law.

$$G^{\frac{1}{2}}B = \text{constant} . \qquad (3\text{-}11)$$

With a single broadbanding circuit

$$G^{\frac{1}{4}}B = \text{constant} \qquad (3\text{-}12)$$

and with a double broadbanding circuit

$$G^{\frac{1}{6}}B \;=\; \text{constant} \tag{3-13}$$

where the constant is approximately the same.

Table 3 shows the effect of signal broadbanding circuits for a gain bandwidth product of 1000.

<u>Table 3</u>

Gain dB	Bandwidth MHz		
	$G^{\frac{1}{2}}B = 1000$	$G^{\frac{1}{4}}B = 1000$	$G^{\frac{1}{6}}B = 1000$
20	100	315	457
16	158	400	540
10	315	560	680

In a similar manner extra tuning can be introduced in the idler circuit and greater bandwidths can be obtained. A particular broadband idler circuit arrangement is achieved by using a double diode amplifier[11] in which two identical GaAs varactor diodes are mounted in opposition in a cavity. The idler current is confined to the loop formed by the two matched varactors. The symmetrical properties of the circuit also fulfil most of the filtering requirements of the paramp. The idler circuit is broadbanded by coupling in an extra resonant circuit of variable frequency and Q by means of a common coupling capacity which is adjusted by using a very simple mechanical tuning control.

Although the negative resistance parametric amplifier discussed thus far is the most common of the parametric devices the 'degenerate' parametric amplifier also finds application. The degenerate parametric amplifier is a negative resistance amplifier in which both signal and idler frequencies are contained within the passband of the amplifier and with th[illegible]al frequency approximately equal to the idler frequency, *ie* the device is [illegible] approximately twice the signal frequency. The gain and bandwidth p[illegible]es of the degenerate paramp are essentially the same as those discussed for the nondegenerate paramp with one exception. Since the idler power falls within the amplifier passband detection of this power in addition to the power at the signal frequency in effect means a 3 dB increase in gain. In a practical degenerate amplifier there are
76 only two discrete circuits, the signal/idler circuit and the pump circuit.

The particular property of the degenerate paramp is that it has two responses, one at the signal frequency and the other at the idler frequency. Consequently it does not find great application in single response communication receivers. A comparison of the noise performance of the paramp, the upconverter and the degenerate paramp shows that the double response degenerate paramp is best and single response degenerate paramp worst. The double response degenerate paramp thus finds application in radiometry and, in particular, in the radio astronomy field.

The negative resistance paramp is the parametric device most used in communication systems. Table 4 shows a selection of the currently available or under development, paramps.

Table 4

Currently available Parametric Amplifiers

Centre frequency GHz	Noise temperature K	Type	Bandwidth MHz	Application
0.5	7	Cryogenic upconverter	380	Radio astron
1.4	35	Uncooled	30	Satcomm
2.25	140	Uncooled integrated circuit	80	Satcomm
2.2	75	Uncooled	20	Tropo systems
3.5	150	Uncooled	140	Space qualified NASA
4	20	Cryogenic	500	Intelsat term
5	55	Uncooled degenerate	10	Radio astron
7.275	180	Uncooled integrated circuit	100	Military satcomm
7.5	80	Uncooled	500	Military satcomm
7.5	140-160	Uncooled	50	Military satcomm
9.5	200	Miniature uncooled double diode design	up to 500	Airborne radar
11/12	165	Peltier cooled double diode	up to 500	Satcomm
12	50	Cryogenic	500	Satcomm
14	290	Uncooled	500	Space qualified ESA
19	115	Cryogenic	600	Satcomm
22	100	Cryogenic	1 GHz	Development
30	590	Uncooled	300	Satcomm/dev
36	410	Uncooled	100	Satcomm/dev
70	950	Uncooled	670	Radio astron
94	1000	Uncooled	2 GHz	Experimental

The cyrogenically cooled paramps listed in Table 4 require expensive refrigerators and compressors and are therefore only suitable for fixed ground stations. The circulator in these paramps is also cooled and it is necessary that the circulator ferrite material should be chosen to give the optimum characteristics at the required operating temperatures. Peltier cooled paramps are thermoelectrically cooled usually down to between $0^{o}C$ and $-40^{o}C$. The low noise performance of these paramps is achieved by operating with a high pump frequency.

The noise temperatures given in Table 4 include circulator losses. The effect of losses between the antenna and the front end amplifier is dealt with in section 2.3.3(c). It is evident that the circulator insertion loss must be kept to a minimum. At room temperature a 0.5 dB circulator insertion loss will increase a paramp noise temperature by approximately 30 K. If the 0.5 dB insertion loss circulator is cooled to say liquid nitrogen temperatures along with the paramp then the increase in paramp noise temperature will only be about 10 K. At an ambient of 20 K the increase in noise temperature will be about 4 K. Room temperature paramps must use circulators with the minimum insertion loss possible. However, insertion loss is not the only criterion by which circulators are selected. The isolation between ports is of great importance. Fig 14 shows a schematic of a negative resistance parametric amplifier coupled to an antenna via a four port circulator. Ideally only a three port circulator is required to separate the input and output signals. However a four port circulator gives better isolation. Basically any amplified signal power reflected from the load is dumped in the matched termination which increases the isolation between the antenna and the load. Isolation between ports two and one must be good since amplified signals may travel back towards the antenna to be reradiated or, if the antenna match is not good, reflected back into the amplifier. Reflections from the load and the matched termination could similarly find their way back to the amplifier input. Such effects lead to gain ripple across the passband and an increase in the paramp noise temperature. If the paramp is at room temperature T_0 the overall effective input noise temperature, including the circulator is given by

$$T_e = T_{ma} + (L - 1)T_0 + L\left[T_{\ell p} + T_{tp} + T_p + \frac{T_r}{g}\right] \qquad (3\text{-}14)$$

where T_{ma} is the effective noise temperature at the antenna terminals due to
076 noise emitted by the matched termination and reflected at the antenna because of the finite antenna VSWR

$T_{\ell p}$ is the effective noise temperature at the input of the paramp due to load (receiver) noise transmitted directly through the circulator due to finite isolation between ports three and two

T_{tp} is the effective noise temperature due to the matched termination noise and the finite isolation between ports four and two

T_p is the effective input noise temperature of the paramp

T_r is the effective input noise temperature of the receiver including all losses between ports two and three

g is the gain of the paramp

L is the total loss between the antenna terminals and the input of the paramp.

In well designed circulators the isolation between ports three and two and four and two can exceed 30 dB so that $T_{\ell p}$ and T_{tp} are about 1 K which is negligible compared with T_p. If the antenna VSWR is less than 1.1, T_{ma} will also be about 1 K if the matched termination is at room temperature. Effective isolation can be improved by using a five port circulator (two three port circulators in cascade) but only at the expense of insertion loss.

Circulators are constructed in waveguide or stripline with the junctions loaded with ferrite discs or wedges. The ferrite is magnetically biased usually with small permanent magnets. Efficient, or low loss circulators, require low loss ferrite material and careful junction matching. Room temperature circulators are available at all frequencies up to 40 GHz. Waveguide or stripline is used up to 18 GHz with stripline predominant below 4 GHz and waveguide from 18 GHz upwards. Practical MIC circulators have demonstrated less than 0.2 dB insertion loss at 4 GHz with greater than 40 dB isolation. With a paramp stage gain of 12 dB the isolation between the paramp and the antenna is about 30 dB which results in a negligible increase in the total noise temperature. The 94 GHz paramp referred to in Table 4 utilises a 94 GHz five port waveguide circulator providing 0.5 to 0.75 dB insertion loss and greater than 20 dB isolation over a 4 GHz bandwidth.

The last 10-15 years has seen considerable development of the paramp and it is currently the key front end amplifier for satellite communication receivers. It is unlikely to be replaced in the fixed ground station where the best achievable noise performance is required. Such installations could support cyrogenic cooling plants. Some paramp developments have however been aimed at the mobile

market. The miniature and integrated circuit paramps referred to in Table 4 have been used for airborne and space applications. For example the 14 GHz paramp in Table 4 was specifically developed for use in the ESA OTS and ECS satellites. The paramp was designed to be light and rugged. All the microwave components of the amplifier and of the pump generator were designed using evanescent mode waveguides[12]. The pump was a 2.5 GHz transistor oscillator followed by two quadruplers. The 7.25 GHz integrated circuit paramp referred to in Table 4 was constructed entirely from thin film microwave integrated circuits, except the Gunn pump oscillator, which results in a compact, reliable design weighing only 200 gram. It is in this type of application, however, that the paramp is facing stiff competition from the GaAs FET amplifier. The paramp requires a source of pump power which must be accurately stabilised and it must operate with a circulator. Ideally some form of temperature stabilisation, such as Peltier cooling, should be used to ensure stability of the varactor induced paramp noise temperature. Operation at lower noise temperatures when cyrogenic cooling is not used required pump frequencies at least 5× the signal frequency which leads to pump procurement problems as the signal frequency increases. That such problems can be overcome is exemplified by the development by NASA of a prototype single stage paramp to operate in a space environment. A gain of 18 dB, noise temperature of 45 K over the band 2.2-2.3 GHz were achieved in a package weighing only 600 gram. This particular noise temperature at 2.2 GHz cannot be bettered at the present time by a GaAs FET amplifier.

3.5 Transistor amplifiers

Significant recent advances in the fabrication of Si bipolar transistors and, more importantly, GaAs field effect transistors have made transistor amplifiers the fastest growing development area in the low noise front end field. At frequencies 1-4 GHz the bipolar amplifier and the GaAs FET amplifier compete directly. At frequencies above 4 GHz the FET is the preferred device having a gain and low noise capability the bipolar cannot match. State of the art developments in transistor amplifiers have led to noise performance second only to the maser and paramp although potentially over a more restricted frequency range. The advantage of the transistor amplifier arises from the fact that when compared to the negative resistance devices, *ie* the maser, TDA and the paramp it is essentially unilateral and can therefore operate without a circulator. Octave bandwidths are achievable making the amplifier more comparable with the LNTWT. The bipolar amplifier and the FET amplifier are discussed separately.

3.5.1 Bipolar transistor amplifiers

Recent advances in the performance of bipolar transistors have generally been made possible by improvements in the device fabrication technology resulting in smaller geometries and reduced parasitics. Small signal Si bipolars are now fast approaching their ultimate performance, *ie* performance limited by the constants of the semiconductor material rather than fabrication limitations. The upper frequency limit of Si bipolars is limited by the ratio of electron mobility to emitter width. The device emitter width is the smallest of the device lateral dimensions and therefore determines, together with photolithographical limitations, the geometry of the device. Consequently the emitter width determines device capacitances and together with electron mobility, device spreading resistances which ultimately limit the bipolar transistor frequency response.

Microwave bipolar transistors are fabricated predominantly from silicon. The reason is technology. Both Ge and GaAs have higher electron mobilities than Si. Theoretical cut-off frequencies of microwave bipolars fabricated with 1μm emitter widths from Ge, Si and GaAs are respectively 10.4 GHz, 8.6 GHz and 18.5 GHz. The GaAs device is evidently superior in theory. Practically it is extremely difficult to manufacture a GaAs bipolar transistor and consequently there is little work in this field. The simpler geometry GaAs FET transistor is a more readily achievable goal for the GaAs technologist. The Ge device is similarly not widely pursued. Fabrication difficulties render the device unattractive. Silicon on the other hand has an advanced technology, the planar process. Significant recent advances in silicon technology are in the production of abrupt emitter impurity profiles by using arsenic instead of phosphorous, the use of ion implantation methods to control the base impurity profile and the introduction of electron beam photolithography. These techniques have led to the production of lower noise, higher gain and f_T devices. Families of Si npn microwave transistors can be produced at reasonable cost and excellent device parameter repeatability within a production batch is maintained. Some improvements are still possible with electron beam photolithography promising emitter widths of 0.5 μm or even 0.25 μm. Operation of the Si bipolar up to about 8 GHz as an amplifier would seem just feasible but unlikely, as the device noise level increases rapidly with increasing frequency near the device f_T.

The primary noise mechanisms within the microwave bipolar transistor are shot noise and thermal noise. The thermal noise arises predominantly from the base series resistance, the reduction of which has led to lower noise devices, and shot noise arises from the dc current flow through the device. The increase

in base recombination as the device internal gain falls off as f_T is approached causes the noise to increase with frequency.

Fig 15 shows the dependence of noise on frequency for the current state of the art in 1 μm emitter commercially available low noise bipolar transistors. The Si bipolar has the advantage of a long pedigree and its reliability is well established. An MTTF of approximately 10^{10} hours for small signal operation is typical.

The low noise performance of bipolar transistor amplifiers is necessarily worse than that achievable with individual transistors and, as may be expected, the noise performance of octave bandwidth amplifiers is worse than that of narrow bandwidth amplifiers. Table 5 lists some of the currently commercially available low noise bipolar transistor amplifiers.

<u>Table 5</u>

<u>Available low noise bipolar transistor amplifiers</u>

Frequency GHz	Noise temperature K	Gain dB
0.005-0.5	290	28 min
0.001-1.0	360	27 min
1.2-1.4	200	25 min
1.4-2.3	290	25 min
1.0-2.0	290	25 min
2.0-4.0	450	25 min
3.7-4.2	590	8.0 min
5.9-6.4	590	7.0 min

The intercept point for these amplifiers is typically +15 to +20 dBm. Very narrow bandwidth amplifiers could approach the individual transistor noise temperature. The amplifiers can be constructed using MIC techniques which result in generally low cost, if sufficient numbers are produced, particularly as bipolar amplifiers can be made untuneable. The excellent control over the transistor S parameters within manufactured batches means that stable amplifiers using fixed untuned designs can be realised.

Where the bipolar amplifier competes directly with the FET amplifier, *ie* at frequencies below 3-4 GHz the bipolar amplifier may be preferred at least from the amplifier designers point of view. The impedance looking into the bipolar at frequencies up to 4 GHz is relatively low. It is therefore easier to match into
76 the device over a broadband than the FET and the design lends itself easily to production in MIC format.

A bipolar amplifier stage will present an input mismatch when designed for minimum noise and an output mismatch to ensure stability. An isolator could therefore be used between the antenna and a front end low noise bipolar amplifier to improve the VSWR. An alternative approach is to use balanced amplifiers[13]. The balanced amplifier module is shown in Fig 16.

The input signal is fed to the two identical amplifier stages via a 3 dB hybrid coupler and any mismatch is absorbed by the 50 Ω load. Similarly any output mismatch is absorbed by the output coupler 50 Ω load. The hybrid-amplifiers-hybrid module forms a cascadable gain module which enables the amplifier optimum noise source impedance to be provided without input VSWR degradation. However some of the noise temperature improvement is lost due to the insertion loss of the input hybrid coupler. The disadvantage of the circuit is increased complexity and therefore cost and possibly reduced reliability. However on this last point it would be argued that the amplifier stage reliability has been improved since if one amplifier should fail then the module will continue to function but with a reduced performance, *ie* the balanced module has a degree of redundancy built it, a factor which may be important for space or mobile receiver application. Table 6 lists a small selection of currently available cascadable amplifiers.

Table 6

Available cascadable amplifiers

Frequency GHz	Noise temperature K	Gain dB
0.005-1	225	16
0.001-1.2	350	24
0.001-2.0	350	12
1.7-2.3	615	8

The cascadable amplifiers are produced using photolithographically defined thin film lumped elements and can therefore be produced in great numbers ensuring low costs.

The design of both single-ended and balanced designs is well understood and by the judicious choice of transistor it is possible to design bipolar amplifiers with lower noise temperatures than those indicated in Tables 5 and 6, in fact within 5-20 K of the transistor noise temperature.

3.5.2 Field-effect transistor amplifiers

Like the microwave Si bipolar transistor the microwave GaAs FET has improved significantly in the last few years. Unlike the bipolar however, the FET still has considerable development potential left and substantial progress in the near future can still be expected[14]. The FET transistor is basically the device described by Shockley in 1952. It does not, like the bipolar, depend on the transfer of minority carriers across pn junctions. It is rather analogous to the vacuum triode in which the flow of electrons between the cathode (source) and the anode (drain) is controlled by the bias on the grid (gate) electrode. Fig 17 shows a schematic of the metal semiconductor FET the MESFET, an FET which is only one, albeit possibly the most important one for microwave application, of the several FET structures. The conductivity of the channel beneath the gate is modulated by applying a negative voltage to the gate whilst a positive voltage is applied to the drain. The most important geometrical parameter in the MESFET is the gate length. Decreasing the gate length reduces the gate source capacitance, which increases the device transconductance, and reduces the transit time of electrons between the source and drain which improves the device frequency performance. In fact for short gate length devices the MESFET f_T is proportional to 1/gate length. Thus the highest frequency devices are those with the shortest gate length. Of course the mobility and saturated drift velocity of electrons in the channel ultimately limits the device f_T. This is why GaAs is chosen rather than silicon for microwave MESFETS. The mobility of electrons in GaAs is approximately 6× that of electrons in silicon and the saturated drift velocity approximately twice that of silicon. The GaAs MESFET is fabricated by growing an epitaxial n type GaAs layer onto a semi-insulating GaAs substrate. The source and drain are metal ohmic contacts and the gate a metal Schottky barrier. Gate lengths are limited by the photolithographic techniques used to define the gate metallisation. Like the bipolar a minimum dimension of 1 μm is achievable using conventional masks. Electron beam or X-ray photolithography promises gate lengths of about an order of magnitude less although at present 0.5 μm can be achieved. The recent significant improvements in MESFET performance have been brought about by the development of the sophisticated materials technology that enabled high quality low conductivity GaAs epitaxial layers to be grown on semi-insulating substrates. There are still some problems however with this technology at the present time, and material repeatibility cannot always be guaranteed. Apart from
076 GaAs, interest in Indium Phosphide is developing. InP has a 50% higher saturated drift velocity than GaAs so that a higher f_T would be expected from an

InP FET. Comparison of the high frequency performance of 1 μm gate length GaAs FET and InP FET yields f_Ts of 25 GHz and 32 GHz respectively[15]. However the gain and noise performance of InP FET appear at this stage of development to be slightly worse than the GaAs FET. Since it takes 5-10 years for a new materials technology to become established it is unlikely that InP will supplant GaAs as the major FET material in the near future as there is still considerable development potential left in GaAs FET.

The noise mechanisms that contribute to the noise performance of FETs are complex and are still to some extent the subject of discussion and research. There are several contributions but it is apparent that a major contribution is resistive, *ie* thermal noise arising in the channel. The FET noise temperature increases with frequency near the device f_T as its transconductance decreases with frequency. Fig 18 shows the state of the art in low noise GaAs FET devices where achievable noise temperatures are shown as a function of frequency. Since thermal noise is a major component of the FET noise temperature operation at reduced ambient temperatures is possible with a reduction in the FET noise temperature.

GaAs FETs can also be fabricated with dual gates[16] in which the second gate acts as a gain control electrode. A slight penalty is paid in noise temperature but higher gain FETs can be realised. Typical figures are 280 K and 10 dB associated gain at 12 GHz for 0.5 μm single gate FETs and 415 K and 13.2 dB associated gain at the same frequency for dual gate FETs. The gain control capability of dual gate FET could usefully find application in remote mobile transponders. There is no parallel capability in the bipolar transistor.

The reliability of the GaAs FET is not yet as good as the bipolar. MTTFs of about 10^7 hours are predicted for small signal FETs. Reliability studies are continuing and some improvement in the MTTF of GaAs FETs has been noted by passivating the device surface with a thin coating of polycrystalline GaAs.

Two other forms of FET deserve mention: they are the junction field effect transistor, the JFET and the insulated gate field effect transistor, the IGFET. In the JFET the Schottky barrier gate of the MESFET is replaced by a diffused or implanted pn junction, the reverse biasing of which controls the channel conductance. Developments have not been as rapid as with the MESFET for technology reasons. It is more difficult to diffuse or implant a closely defined p region in the n epitaxial layer than it is to deposit a metal Schottky barrier. It is possible that the JFET may produce a lower noise device since the channel 076
will be deeper within the epitaxial layer and therefore further from the disordering effects of the surface. The IGFETs are more of interest for power

amplification at microwave frequencies. They can operate with large active gate voltages which take them from enhancement mode to depletion mode operation. These devices have been realised predominantly on silicon since it is relatively easy to insulate the gate from the substrate using silicon dioxide. Some work is underway to produce insulating layers on GaAs although results so far are poor compared with SiO_2 on Si. Low noise 1 μm gate length IGFETs in both single and dual gate configurations are produced for use at frequencies up to 1 GHz.

The FET is used in small signal amplifier circuits from about 2 GHz upwards. Below 2 GHz the very high impedance seen looking into the device makes impedance matching difficult particularly over a broadband. Both single ended and balanced designs are realised by a very large number of manufacturers with the UK capability among the state of the art leaders. Table 7 lists a selection of the currently available GaAs FET amplifiers.

Table 7

Available GaAs FET amplifiers

Frequency GHz	Noise temperature K	Gain dB
2.0-4.0	280	10 min
2.7-3.1	185	14 min
3.1-3.5	220	10 min
3.7-4.2	125	40 min
4.0-8.0	430	12
8.0-12.0	615	9
11.4-12.6	720 uncooled 90 cooled to 40 K	8
13.75-14.25	615	8
12.0-18.0	850	5.5

The intercept point of these amplifiers is typically +15 dBm. The amplifiers are constructed predominantly using microstrip techniques. At the lower frequencies encapsulated devices are used but at higher frequencies optimum performance is achieved in amplifier designs that utilise the unpackaged FET chip. Monolithic amplifiers can also be obtained. Some recent work[17] has demonstrated the feasibility of producing GaAs FET amplifiers on a GaAs substrate, *ie* the input and output impedance matching elements are realised using thin film lumped elements which are actually fabricated on the same GaAs substrate as the FET, hence the monolithic amplifier. Small size and high reliability as well as
076 potentially lower costs can be achieved in this way. As with bipolar transistors the source impedance that results in minimum noise produces an input mismatch.

Amplifiers are therefore produced with integral isolators or in balanced formats in order to yield respectable input and output VSWRs. The parameter repeatability of the FET is not as good as the bipolar. Consequently most designs are made tunable to some degree.

Table 7 shows the results of cooling an FET amplifier. The particular set of results[18] is illustrated in Fig 19 in which the gain and noise temperature of a three stage 12 GHz FET amplifier is shown as a function of frequency for different values of ambient temperature. The decrease in the noise temperature is expected if thermal noise is a major component of the FET noise mechanisms. The increase in gain is due to the increase in the device transconductance with decreasing temperature. The amplifier was designed for operation at 90 K in the 11.7-12.2 GHz satellite communication band. With improved matching networks, *ie* less lossy, it was estimated that the noise temperature at a 90 K ambient could be as low as 92 K compared with 148 K actually measured.

The amplifier was a microstrip on sapphire design in which the GaAs FET were utilised in chip form, *ie* unpackaged. The chips were mounted directly on the substrate ground plane which acted as a heat sink. If cooling down to 90 K or lower is not possible in a particular system application it is evident from Fig 19 that a modest amount of cooling, possible using Peltier plates, results in a considerable reduction in the amplifier noise temperature. Cooling, using Peltier plates and/or radiation coolers, is possible in satellite applications in which GaAs FET amplifiers are used as low noise front ends in transponders. Staggered cooling, *ie* cooling of the first stage of a front end amplifier cascade, may be all that is necessary to significantly reduce the front end receiver noise temperature.

The comparison of the cooled GaAs FET amplifier with cooled and uncooled paramps may be made by reference to Table 4. At 12 GHz a Peltier cooled (~270 K ambient) paramp has a noise temperature of 160 K. When cooled to liquid nitrogen temperature, *ie* 70 K the paramp noise temperature is $<$50 K. At 90 K the GaAs FET amplifier noise temperature will be 92 K. In terms of noise performance the paramp is thus still superior and an even lower noise temperature can be achieved by pumping the paramp at a higher frequency. The GaAs FET amplifier is still left with the residual non-resistive noise components - indeed at an ambient of 40 K the amplifier referred to in Fig 19 showed no further reduction in noise. However the comparison is not made on noise performance alone. At the 1 dB gain
compression point the GaAs FET amplifier typically delivers +10 dBm of output 076
power compared with typically -15 dBm for the paramp. Larger bandwidths are

available with the GaAs FET amplifier and gain stability is achieved without the special care required for the paramp. Cooled GaAs FET amplifiers with their simple circuitry and potential low cost will therefore compete strongly with paramps at frequencies up to 18 GHz.

3.6 Mixers

The mixer is the basic component of the heterodyne receiver. The efficiency with which it downconverts the incoming RF signal to the chosen IF contributes to the overall noise temperature of the receiver. The degree to which it contributes depends, as is shown in equation (2-12), upon whether there is RF amplification between the antenna and the mixer. Before the advent of the Schottky barrier diode mixer, point contact diodes were used which were both noisy and lossy. The Schottky diode is however capable of producing near theoretical noise temperatures and conversion losses[19]. Consequently the Schottky mixer diode has made possible receiver mixer front ends which can compete with RF amplifier front ends at some frequencies. The Schottky diode is simply a metal-semiconductor junction. Its application as a mixer arises from its nonlinear resistance when forward biased. Good high frequency and noise performance requires low minority carrier injection, low junction capacitance and low series resistance. The minority carrier injection requirement precludes the use of the *semiconductor* pn junction. GaAs is used rather than Si for the highest performance Schottky diodes as the series resistances are lower than in the comparable Si diode. Cut-off frequencies well in excess of 500 GHz can be achieved with GaAs Schottky diodes.

The requirement for isolation between the RF input, the local oscillator, and the IF output determines, together with the conversion loss or gain between the RF input and the IF output, the manner in which the mixer diode is used in circuit. Mixer circuits can be classified as single-ended (or unbalanced) mixers, which use a single diode, double-ended (or balanced) mixers which use two diodes, and double balanced mixers which use four diodes. Fig 20 shows schematics of the three basic mixer types.

The single-ended or unbalanced mixer is the simplest mixer but its performance is limited. Isolation between the LO, RF and IF ports is inherently poor and LO am noise contributes to the noise performance of the circuit. However the single-ended mixer can find application at frequencies where the hybrids used in more complex mixers become lossy, thus degrading the mixer noise performance. Port isolation can be improved if the IF frequency is chosen to suit the rejection
076 properties of practical filters. The balanced mixer shown in Fig 20b improves the isolation between the LO and the RF and IF ports. If a 90° hybrid is used

the LO to RF isolation is dependent upon the diode impedance match since LO energy not absorbed by the diode is returned to the RF port. Since diode impedance is a function of LO power this isolation is dependent on LO drive. This isolation is improved if a 180° hybrid is used since the LO/RF isolation now becomes a function of the diode balance rather than the diode impedance. LO drive will not affect this isolation provided the diodes continue to track each other. The symmetry of the circuit results in the cancellation of LO am noise and the production of fewer mixing products than the single-ended mixer. However, there is no isolation between the RF and IF ports. The third basic mixer circuit is shown in Fig 20c&d. The double balanced mixer offers, in addition to the properties of the single balanced mixer, isolation between all three ports thus considerably easing the filter problem. The circuit is more complex however and requires more LO power. Conversion loss is also increased.

Ideally the minimum conversion loss that can be achieved with any of these circuits is 0 dB but conversion gain is not possible with the nonlinear element purely resistive. To approach this minimum loss it is necessary that the mixer diode should be terminated in tuned circuits that support only the RF signal and IF output frequencies. Other modulation products must see lossless terminations at their respective frequencies. It has been shown however that the conversion loss of any broadband mixer is constrained theoretically to be at least 3 dB [20] since it is supposed that the image frequency ($2f_{LO} - f_{sig}$), the most important of the modulation products other than the wanted IF frequency, is sufficiently close to the signal frequency to dissipate power in the signal circuit. However if the signal and image frequencies can be separated and the image termination of the mixer made lossless (*ie* purely reactive, short circuit or open circuit) then the theoretical minimum conversion loss is reduced to 0 dB. In practice it is not possible to provide a lossless termination for the image power. This has prompted the attempt to realise practical mixer circuits in which the image frequency modulation product is recovered and not wastefully dissipated in a resistive termination. The usual circuit technique is to reflect the image power back into the mixer for reconversion to the wanted IF. Half of this power is, by reciprocity, converted to the original signal power to be dissipated in the signal load admittance. Thus image recovery ideally results in a conversion loss of 1.5 dB. Circuit losses and mismatch would make this figure slightly higher. Practical mixers have not yet reached these ideal conversion losses and the technique of image recovery is still an active research and development field. At the present time improvements in mixer performance appear to depend on improved 076
circuit techniques, rather than on the mixer diodes.

In practice there are two circuit techniques that can result in image recovery. One is to use a selective filter in the signal line to separate the signal and image frequencies. The suitable inclusion of the filter, which passes the signal but rejects the image, at the correct electrical distance from the diode will cause the reflected image to be correctly phased at the diode so that the wanted IF it produces will be in phase with the IF produced by the original signal. If the IF frequency is however low, *ie* the image and signal frequencies are close, then the filter will require a correspondingly high Q . This technique is therefore used mostly for high IF frequency mixers. The second method is to connect more than one mixer in such a manner that image recovery can occur by virtue of the symmetry of the circuit. Fig 21 shows a schematic of a commercially available image recovery mixer which utilises more than one mixer[21]. The feature of the circuit is the direct RF connection of both double balanced mixers. This circuit arrangement provides a reactive image termination with the common signal junction appearing as a short circuit at the image frequency. Additionally the image signal generated by either mixer acts as an input to the other mixer which produces wanted IF in addition to the main RF to IF conversion. Table 8 compares the practical performance of the basic commercially available mixer types.

Table 8

Comparison of mixer types

Mixer type	No. of diodes	VSWR	Conversion loss	Intercept point**	LO/RF isolation
Single balanced 90° hybrid	2	1.3:1	5-7 dB	+15 dBm	10 dB
Single balanced 180° hybrid	2	2.5:1	7-9 dB	+15 dBm	20 dB
Double balanced	4	2.5:1	7-9 dB	+20 dBm	>30 dB
Image recovery	8*	2.0:1	3-4 dB	+15 dBm	30 dB

* Uses two double balanced mixers

** Typically 6-9 dB above LO power

Intermodulation distortion in mixers is measured in similar way to amplifier distortion. Two RF signals are applied to the mixer and the third order intercept
076 point is measured. Intermodulation distortion basically arises when the mixer conversion loss departs from linearity. When the wanted IF output cannot follow

the RF input linearily, the 1 dB compression point is defined when the mixer conversion loss increases by 1 dB.

Mixer noise is closely related to conversion loss which is a function of the circuit components and arrangement. The noise generated within the mixer diodes themselves is also a contributory mechanism. Mixer noise is usually specified as a noise figure in dB rather than in equivalent noise temperature terms due to its close relationship to conversion loss and is usually written

$$N_m(\mathrm{dB}) = 10 \log_{10}(L_c t_m) \qquad (3\text{-}15)$$

where L_c is the mixer conversion loss and t_m the effective mixer noise temperature ratio. L_c arises basically from the mixer diode conversion loss but also includes contributions from mismatch and losses arising in other components within the mixer circuit. In simple terms t_m is the ratio of the mixer diode noise temperature to the standard temperature of 290 K. The basic noise contributions within the mixer diode are due to shot noise and thermal noise and in modern Schottky barrier diodes t_m can be as low as 0.85. t_m is dependent on the quiescent current through the diode and the IF frequency. For very low IFs flicker noise contributions from the mixer diode can make $t_m > 1$. A typical value of t_m is however near unity, consequently the mixer noise figure is approximately the same as its conversion loss and quite often less.

Since mixers normally work into an IF amplifier and the mixer impedance termination at its IF output frequency must be closely controlled, IF amplifiers are often provided with, or integrated with, the mixer. Mixer noise figures therefore usually include the contribution from the following IF amplifier. Tables 9, 10 and 11 list some of the currently commercially available single balanced, double balanced and image recovery mixers respectively.

180° hybrid single balanced mixers are available to cover the same ranges. Noise figures are 0.5 dB greater and IF bandwidths are approximately 50% of those shown below.

Table 9

Available single balanced mixers

RF frequency GHz	NF (SSB) (max) dB	Eq noise temperature K	Type	IF bandwidth MHz
0.5-1.0	7 (inc 1.5 dB IF amplifier contribution)	1160	Coax, 90^{o} hybrid	DC - 150
1.0-2.0	7 (inc 1.5 dB IF amplifier contribution)	1160	" " "	DC - 225
2.0-4.0	7 (inc 1.5 dB IF amplifier contribution)	1160	" " "	DC - 600
4.0-8.0	7.5 (inc 1.5 dB IF amplifier contribution)	1340	" " "	DC - 900
8.0-12.4	7.5 (inc 1.5 dB IF amplifier contribution)	1340	" " "	DC - 1200
12.4-18.00	8.0 (inc 1.5 dB IF amplifier contribution)	1540	" " "	DC - 1400
26.0-40.0	9.0 (inc 1.5 dB IF amplifier contribution)	2000	Waveguide 22	3 MHz at 30 MHz

Table 10

Available double balanced mixers

RF frequency GHz	NF (SSB) (max) dB	Eq noise temperature K	IF bandwidth MHz
1.1-2.7	7.0 (inc 1.5 dB IF amplifier contribution)	1160	DC - 350
2.5-9.5	8.0 (inc 1.5 dB IF amplifier contribution)	1540	DC - 500
5-11	8.5 (inc 1.5 dB IF amplifier contribution)	1780	DC - 900
8.0-18.0	9.5 (inc 1.5 dB IF amplifier contribution)	2280	DC - 4000

Table 11

Available image recovery mixers

RF frequency GHz	NF (SSB) (max) dB	Eq noise temperature K	IF bandwidth MHz
1.04-1.14	3.8 (inc 1.0 dB IF amplifier contribution)	410	25-35 50-70
1.0-2.0	4.5 (inc 1.0 dB IF amplifier contribution)	515	25-35 50-70
3.7-4.2	4.3 (inc 1.0 dB IF amplifier contribution)	490	25-35 50-70
8.5-9.0	5.3 (inc 1.0 dB IF amplifier contribution)	700	25-35 50-70

Several laboratories have reported slightly lower noise figures and higher frequency image recovery mixers. Developments may be expected in the near future in both circuit techniques to improve mixer conversion loss, since the theoretical minimum has not yet been reached, and in improved higher cut-off frequency mixer diodes. In particular the beam lead GaAs Schottky diode holds the greatest promise for cut-off frequencies in excess of 500 GHz.

The matched image mixers although of higher noise figure than image recovery mixers offer wide RF bandwidths with greater than an octave commonplace. Image recovery mixers which rely on image filtering can achieve relatively narrow RF bandwidths while those utilising the circuit arrangement of Fig 21 offer octave bandwidths together with a low conversion loss. For single response receiver applications the low conversion loss and hence low noise image recovery mixers are an obvious candidate since those mixers of Fig 21 are also image rejection mixers, *ie* they have usually better than 20-25 dB rejection of RF input signals appearing at the image frequency. The single and double balanced mixers of Tables 9 and 10 respectively must be used with appropriate filters to suppress the image response and noise input at the image frequency.

The mixers described thus far are all constructed using microwave integrated circuit construction techniques and beamlead Si or GaAs Schottky diodes. They are therefore rugged, reliable and of small size typically 20 cm^3 or smaller. The pairs and quads of Schottky diodes are fabricated on the same Si or GaAs substrate. The beamleaded substrate may then be mounted into a microstrip circuit arrangement. The use of thin film techniques can ensure low cost volume production. 07[

Most of the mixers described require LO power of 0-10 dBm. Some are available with a dc bias option on the mixer diodes enabling the use of low power (-10 dBm) LO sources. Typical dc power requirements for reduced LO power operation are 12 V and 2 mA.

Fairly recently the GaAs FET has been investigated for potential application as a mixer. The GaAs FET has a Schottky barrier between the gate and source. By using this Schottky barrier as a mixer it was anticipated that the IF frequency produced by the mixer would be simultaneously amplified by the FET. Using FETs designed for amplifier application it was shown that this was the case[22]. However, the conversion loss of the gate source Schottky barrier was found to be about 15 dB which is considerably larger than the conventional Schottky barrier diode mixer conversion loss. Including the IF amplification by the FET an overall conversion gain of 6 dB was achieved. The noise figure was high, close to the mixer conversion loss of 15 dB. Experiments were carried out with a signal frequency of 10.8 GHz and an IF of 1.7 GHz. The experiments showed promise however and point to the requirements for FETs designed for mixer application. The goal of such developments is the integrated front end in which GaAs FETs will form the nucleus of RF amplifier, mixer, oscillator and IF amplifier circuits. Low cost and high performance are thought achievable by the fabrication of all the GaAs FETs on one substrate or alternatively by the integration of all the component circuitry on a lower cost substrate to be encapsulated in a single package.

Such developments as these together with the expected advances in image recovery mixers and the current performance of mixers mean that the mixer front end should be considered at all frequencies along with the RF amplifier front end.

3.7 Front end summary

Fig 22 illustrates the state of the art noise performance of the RF front end components described in section 3 as a function of frequency.

4 DISCUSSION

The preceding section described the choice of receiver front end components available to the system/receiver designer. Obviously the selected front end must meet system requirements but overspecification on noise performance must be avoided on cost grounds. As was pointed out in section 2.2 the downlink is most often the critical path in satellite communication systems. The lowest noise receiver is therefore usually included in the ground station. Low noise receivers may however be required in satellite transponders where the satellite is

illuminated from low power transmitters, *eg* from mobiles such as ships, aircraft or manpacks.

Equation (2-6) gives the carrier-to-noise ratio at the input to a ground based or mobile receiver. The final noise level presented to the receiver demodulator is determined by the uplink, satellite transponder, and downlink noise levels and the total system noise budget must be allocated accordingly. The system carrier-to-noise ratio can simply be written

$$\left(\frac{C}{N}\right)^{-1} = \left(\frac{C}{N_{up}}\right)^{-1} + \left(\frac{C}{N_{sat}}\right)^{-1} + \left(\frac{C}{N_{down}}\right)^{-1} \tag{4-1}$$

where N_{up}, N_{sat} and N_{down} are respectively the uplink, satellite transponder (including intermodulation noise), and downlink noise power respectively. Fig 23 shows the total system C/N as a function of the downlink C/N for different values of the combined uplink and satellite transponder C/N. The graph illustrates the point at which further improvements in the downlink C/N through an increase in the G/T ratio for the receiving station, results in little or no measurable improvement in the system C/N because of the eventual dominance of the satellite or uplink noise. Indeed in some frequency multiplexed systems the intermodulation noise generated by the multicarrier operation of transmitter TWTs or transistor power amplifiers dominates the overall system C/N. In such cases overall system improvement can only be brought about by the reduction in this transmitter induced noise. In those systems where the required receiver G/T ratio is specified the trade-off choice is simply between G , the receiving antenna gain and T the downlink noise temperature. The type of antenna chosen for the receiver however depends essentially on the degree of directivity required. A mobile receiver such as a ship or aircraft requires a receiving antenna which will keep the satellite within its main receiving pattern irrespective of the motion of the mobile. The gain (or directivity) of such an antenna may be as low as 3 dB. A fixed ground station antenna designed to receive from a geostationary satellite can however afford a much higher degree of directivity (or gain) and therefore a much larger diameter antenna consistent with budgetary restrictions and the relative cost advantage, if any, over the alternative of reducing the receiver noise temperature. Figs 7 and 8 show the costs of antennas as a function of antenna diameter and the cost of reducing system noise temperature respectively.

The final system C/N is of course determined by the system application. 076
Table 12 shows typical FM system channel carrier-to-noise density ratios necessary

for the transmission of various types of information of acceptable quality. The C/N is expressed in units of dB - Hz and includes a margin of 4 dB over a receiver threshold of 10 dB. Noise power is defined by equation (2-1) which shows that the noise power is directly proportional to bandwidth. The noise power in Table 12 is simply written in terms of kT per Hz as the information channel bandwidth requirements are different for the various services.

Table 12

Typical FM system channel quality requirements

Service	Approximate C/N system required (per channel)
Colour TV	88 dB - Hz
Monochrome TV	85 dB - Hz
Telephony (CCIR quality)	60 dB - Hz
Telephony (tactical quality)	55 dB - Hz
FSK telegraphy	41 dB - Hz

Given that antenna 'gain' may be determined by the frequency of operation and the directivity required, the receiver G/T ratio will be determined by the receiver noise temperature. Some examples of typical receivers will now be discussed.

4.1 An 11 GHz receiver front end

Several proposed satellite communication systems utilise frequencies within the band 11-12 GHz as a downlink. Such systems as Intelsat V, OTS and ECS will use this frequency for the transmission of television, telephone and data. Receiving antenna diameters of about 20 m may be expected for reception from Intelsat V. The ground station G/T requirement for Intelsat V reception is estimated at between 39 dB/K to 44 dB/K, depending on the ground station geographical location and hence the degree of system margin required for adverse weather conditions. Assuming for illustrative purposes a downlink frequency of 11.5 GHz, a 20m antenna has a gain of approximately 64.5 dB yielding an associated receiver noise temperature of 354 K for a G/T of 39 dB/K and 112 K for a G/T of 44 dB/K.

076 The G/T requirements are interpreted as downlink parameters only. Thus T represents the ground station receiver operating temperature, *ie* the noise

temperature arising from the antenna and receiver contributions only. The 112 K requirement indicates that a cryogenically cooled parametric amplifier will be required. A maser though capable of achieving a lower noise temperature would most likely be rejected on cost grounds. None of the other potential front ends would be suitable on noise performance grounds. A typical receiver layout is shown in Fig 24.

The receiver effective input noise temperature is given by equations (2-13) and (2-14) and $T_{op} = T_a + T_e$. At 11.5 GHz the sky noise contribution is, from Fig 2, approximately 17 K (for a 100% efficient antenna) at 10^o elevation, the worst allowable case. The contribution from T_m, discussed in section 2.3.3(a) is difficult to estimate and varies with antenna design. A figure of about 15 K is reasonable which makes the worst case total antenna contribution T_a 32 K leaving 80 K to be assigned to the receiver for a 112 K operating temperature ground station.

The receiver input losses shown in Fig 24 arise predominantly from the antenna waveguide feeder insertion loss which is approximately 0.15 dB/m at 11.5 GHz. Assuming that the whole of the receiver front end is mounted on a Cassegrain antenna close to the feed horn then the feed waveguide will at worst be 1-2 m long giving an insertion loss of 0.15-0.3 dB. Using equation (2-13) and assuming that the waveguide is at an ambient temperature of 290 K, then for an insertion loss of 0.3 dB the waveguide noise temperature is 20.7 K. A cryogenically cooled two stage paramp may be realised with a noise temperature of 50 K and a gain of 24 dB, and a typical mixer SSB conversion loss is 7.0 dB. Assuming an IF amplifier noise figure of 2.0 dB (noise temperature 170 K) then equation (2-14) yields the receiver effective input noise temperature of 84 K of which 20.7 K is due to the feed waveguide noise. The noise contribution of circuits following the IF amplifier will be insignificant for a high value of IF amplifier gain. The inclusion of a 1 dB insertion loss image rejection filter between the paramp and the mixer will add 3 K to the receiver noise temperature. If sufficient image rejection cannot be achieved with a filter and the image falls within the paramp bandwidth a double downconversion may be necessary to provide the necessary rejection.

The receiver noise temperature exceeds the required 80 K by a few degrees. Reducing the feeder waveguide run to 1 m will reduce its noise contribution to 10 K and bring the receiver into the required specification. It is interesting to note the significant contribution to the receiver noise temperature that the feed 076
waveguide makes. The obvious solution is to minimise the feed length although

this may not be practical and will depend on the type of antenna used. Alternatively the feed ambient temperature could be reduced by cooling but this is unlikely to be practical.

Mounting the front end package of Fig 24 immediately behind or close to the main reflector of any antenna introduces formidable engineering problems particularly as the parametric amplifier must be cryogenically cooled. A recently reported new type of Cassegrain antenna feed, the 'beam waveguide feed'[23] which makes use of a series of reflectors to transmit the radiation collected by the antenna to a ground situated RF front end, would appear to overcome the major difficulty of mounting a cryogenically cooled paramp behind the reflector. The RF front end can be situated within a controlled environment at ground level. The insertion loss of the beam waveguide feed was 0.25 dB at 4 GHz, equivalent to approximately 7 m of waveguide at 4 GHz. The waveguide feed however included a diplexer to separate the transmit band from the receive band.

In systems such as these, the antenna feed will continue to contribute a significant proportion of the receiver noise temperature. The receiver operating noise temperature can only be reduced by choosing lower noise components in the system shown in Fig 24. The alternative in this particular system is to use a larger diameter antenna. For a receiver operating noise temperature of 120 K, including 88 K for the effective noise temperature of the receiver, the required G/T of 44 dB/K can be met with a fractionally larger antenna. The antenna/receiver trade-off is ultimately, in this case, a cost effectiveness exercise.

4.2 A 1.5 GHz receiver front end

The 1.5 GHz downlink band is utilised by many satellites that offer communication services and navigational and meteorological information to mobiles, including Marots, Marisat and Navstar. Mobiles such as ships would normally carry antennas of 1.0-1.5 m diameter, typically 1.3 m for reception from Marots or Marisat. The gain of a 1.3m antenna is approximately 24 dB and typically the receiver G/T requirement is -4 dB/K giving a receiver operating noise temperature of 630 K. At 1.5 GHz sky noise is, for 10^{o} antenna elevation worse case, from Fig 2 approximately 13 K assuming a 100% efficient antenna. The contribution from T_m varies between antennas but may be expected to be of the same order as the sky noise. The antenna noise temperature is thus approximately 26 K worst case at 1.5 GHz. Shipboard antennas normally operate from inside radomes for
076 weather protection. There is inevitably loss associated with the use of a radome, typically 0.1-0.2 dB. The effect of radome loss on the receiver noise temperature

may be calculated from equation (2-15) since radome attenuation results in the same effect as attenuation due to rain. A 0.2 dB radome loss will add 12 K to the antenna noise temperature which thus totals 38 K. The receiver noise temperature must therefore be less than 592 K.

The receiver front end will be similar to that shown in Fig 24. The input losses due to the antenna feed will be smaller than the previous receiver example, typically 0.01 dB, due to the ease with which the front end amplifier can be mounted directly behind the main reflector of a small antenna dish. However most mobiles such as ships carry radar and it is often necessary to protect the RF front end of the receiver from radar pulse breakthrough. Consequently, a limiter would be included between the antenna and the RF amplifier to protect the amplifier and successive components. A passive PIN diode limiter operational between 1-2 GHz has typically an insertion loss of 0.4 dB. A diplexer may be included if the antenna is also used for transmission - a typical diplexer insertion loss is also about 0.4 dB. The receiver input losses thus arise from the diplexer and limiter which must both be connected between the antenna and RF amplifier. Application of equations (2-13) and (2-14), assuming a mixer conversion loss of 5.5 dB and conservatively, an IF amplifier noise figure of 2 dB (noise temperature 170 K) yields, for an RF amplifier gain of 30 dB and noise temperature of 288 K, a receiver effective noise temperature of 406 K which is well within the 592 K requirement. The RF amplifier gain and noise temperature specification is typical of commercially available low noise bipolar transistor amplifiers. Fig 25 shows the receiver layout.

In order to meet the receiver noise temperature specification of 592 K the RF amplifier noise temperature cannot exceed 443 K at 30 dB gain. This noise temperature excludes the tunnel diode amplifier, and the mixer from consideration as the first RF front end component. The paramp is obviously not necessary but the GaAs FET amplifier could be considered. The bipolar amplifier would however be more than adequate and would be slightly cheaper than the GaAs FET amplifier.

Mobiles such as aircraft should in principle use antennas with a gain of 3 dB (*ie* hemispheric coverage) in order that reception from a satellite is not lost during aircraft manoeuvres. In practice however some increased degree of directivity may be necessary in order to avoid unwanted interference from other transmitters. Satellites transmitting navigational information in telegraphic format only may require a narrow bandwidth for this purpose, say 100 Hz. In order to maintain a channel quality of approximately 41 dB - Hz a receiver G/T of 076

-26 dB/K is required for a typical application. Thus a 3 dB gain receiving antenna requires a receiver operating temperature of 790 K to meet the system requirement. As before, the antenna noise contribution will be about 26 K which is almost negligible in this case. However since the antenna gain is so low and the aircraft must manoeuvre it is probable that noise will enter the system via the antenna from, for example, the earth which has a ambient temperature of about 300 K. Consequently it is more realistic to include in the antenna noise temperature noise arising from this source together with sky noise. Thus the effective input noise temperature of the receiver must be about 460 K to meet the system specification.

The receiver would be similar to that shown in Fig 25, excluding the diplexer, if the receiver is to receive navigational information only. Protection would be necessary in an aircraft radar environment and an information channel filter, possibly tuneable, included in the receiver after the IF amplifier. Losses within the receiver can also arise from cable runs if the geometry of the aircraft or receive antenna is such that the RF front end amplifier cannot be mounted immediately behind the antenna. Losses from an RF switch, typically a PIN diode switch, must also be considered if the receiver is connected to more than one antenna. A typical receiver layout might be as shown in Fig 26.

Assuming a PIN diode switch insertion loss of 0.6 dB, cable loss at, say, 0.8 dB (for a 3 m run), limiter insertion loss 0.4 dB, mixer conversion loss 5.5 dB and IF amplifier noise temperature 170 K, then from equations (2-13) and (2-14) the RF amplifier noise temperature must not be greater than 205 K with an amplifier gain of 30 dB to meet the system requirement. This amplifier performance could easily be met by a paramp but this would be an expensive overkill. The logical choice is a bipolar transistor amplifier which is capable of meeting this noise performance at ambient over a narrow frequency band of say 20 MHz. The receiver noise temperature includes 58.6 K from the cable loss and 51.6 K from the RF switch. Obviously if these noise contributions can be reduced the noise specification of the RF amplifier can be eased.

4.3 Receivers for 20/30 GHz systems

Satellite communication systems that make use of 20/30 GHz frequencies will no doubt be introduced in the not too distant future. The major advantage of this frequency band is that large bandwidth (~2.5 GHz) systems are thought
076 possible. A drawback however of the use of such high frequencies is that atmospheric absorption increases particularly during rain. Ground station diversity,

in which more than one receiving ground station is used situated 20-30 miles apart, is proposed as a method of overcoming the problem since it is unlikely that both ground stations will be affected by the same rain belt. While system specifications are yet to be defined it is possible to indicate the form of receiver front ends that will find application in these systems.

The satellite transponder front ends, operating at the uplink frequency of 30 GHz, will be specified in a trade-off with the ground transmitter EIRP. At 30 GHz the transponder front end could be either a paramp, a mixer or a TDA, with the possibility, with future development in mind, of a GaAs FET amplifier. The mixer and the TDA have the advantage over the paramp of being able on current performance to meet the complete bandwidth requirement of up to 2.5 GHz. However the noise temperatures of the ambient temperature paramp, mixer and TDA are respectively 400 K (for a bandwidth of 300 MHz), 1000 K (for a projected image recovery mixer) and 1500 K. The paramp noise temperature could be further reduced by Peltier cooling.

A typical 30 GHz satellite transponder receiver might simply be as shown in Fig 27 in which a channel defining filter is shown preceding a broadband paramp. Broadband paramps with bandwidths of approximately 2 GHz should be realisable with further development. It is possible however that more than one transponder will be necessary to cover the fill 2.5 GHz bandwidth. The LO frequencies are chosen to provide a first IF of about 10-15 GHz, and a second IF of 0.5-2.0 GHz. Both IF amplifiers could then be transistor amplifiers, the first IF amplifier using GaAs FETs and the second bipolar transistors.

The weight and power requirements of 30 GHz satellite borne paramps are difficult to estimate but it is reasonable to expect that the weight per complete paramp stage could be kept below 1.0-1.5 kg and dc power drain including pump supplies to below 6 W per stage.

The system noise performance together with the front end dynamic range requirement would necessarily determine whether the paramp, mixer or TDA were chosen as the first RF component. The 20 GHz ground station receiver noise temperature is of course determined by the antenna diameter chosen. At this frequency the RF front end component could be either a paramp, mixer, TDA or a GaAs FET amplifier, assuming the development of the latter in time for the system requirement. The ground station noise temperature must be specified to allow for increased sky and atmosphere noise due to rain attenuation. 076

4.4 Future phased array systems

Communication between relaying satellites and ground or mobile receivers is, as has been previously stated, usually dominated by the satellite-to-ground link, *ie* the downlink. The quality of the received channel information is determined by the satellite EIRP and the ground, or receive, station G/T. Thus far, only the ground station G/T has been considered. Either the receive antenna size and hence gain must be increased or the receive station noise temperature reduced. In some instances however the channel quality cannot be significantly improved due to cost limitations on the low noise receiver, or for some mobiles receive antenna size and pointing constraints. In such cases an increase of transmitter power or the increase in transmitter antenna gain are the only ways in which channel quality can be improved and the number of channels transmitted increased.

It is proposed that some future satellite communications systems should utilise multi-beam phased array antennas in transmission from satellites. The increased directivity, *ie* high gain, achievable by using the phased array antenna means the satellite EIRP is correspondingly higher. Earth coverage is maintained by the generation of a sufficient number of beams each carrying different channels. Fig 28 shows a schematic of the single beam antenna and an example of a multi-beam antenna for a geostationary Maritime satellite operating at *1.5 GHz*.

The phased array antenna is designed to provide 19 beams for earth coverage with an EIRP of 24 dB which is 6.4 dB in excess of the earth coverage single beam reflecting antenna. The G/T of the mobile receiver of section 4.2 could thus be relaxed to -10 dB/K for the same C/N ratio. The receiving antenna could therefore be simplified and/or the receiver noise temperature specification relaxed with a resulting cost saving.

Each element of the array is fed by a linear power amplifier which is fed, via an upconverter, from a beam forming network operating at IF. The beam forming network allocates channels to beams and generates the necessary array feed signals from phased samples of each beam input signal. The disadvantage of the system is the complexity of the beam forming network. It follows that the system is less reliable than the single beam system. However the phased array transmitter is subject to a graceful degradation as the elements and their addressing circuits fail. The phased array can also be used for the satellite receive function if a diplexer, one for each array element, is incorporated between the power amplifiers
076 and the array. A separate receive network to extract the channels will be necessary. The increased satellite receive antenna EIRP will reduce the power transmission requirement from a mobile.

The system is still in the study phase and several alternative multi-beam systems are being examined[24,25]. Basically the alternative systems are fixed position, switchable, *ie* discrete beam positioning, or steerable beams. Whichever phased array system is chosen the specification of the receiver G/T will be relaxed or the communication system capacity will be increased. Either way, multi-beam phased array transmitters and receivers are expected to make a significant impact on future satellite communication systems.

5 CONCLUSIONS

The Report has concentrated on the noise performance of the various types of receiver front end components. Essentially the receiver must be configured such that its effective input noise temperature should be appropriate to the system noise requirement. Those receivers where the receiver downconverter stage will dominate the noise temperature of the receiver require a low noise RF amplifier as a front end with sufficient gain to reduce the downconverter stage noise to an almost negligible level.

For the lowest noise receivers the cryogenically cooled paramp is the first choice since it is more cost effective than the alternative maser. The ambient or Peltier cooled paramp follows as a cheaper second best noise performance RF front end. The GaAs FET amplifier, the Si bipolar amplifier, the tunnel diode amplifier and the image recovery mixer have also been described and must be considered in any system application falling within their frequency capability. In particular, the GaAs FET amplifier and the image recovery mixer may be expected to find increasing application.

Technology improvements in the GaAs materials field are expected to improve the noise and frequency performance of field effect transistors, Schottky barrier diodes and varactor diodes. Consequently the performance of GaAs FET amplifiers, paramps and mixers may be expected to further improve and it is these RF components which are expected to dominate satellite communication receivers. Receivers in which the paramp will provide the first stage of an RF front end amplifier followed by a GaAs FET amplifier and a image recovery mixer would appear the most likely combination for the lowest noise state of the art receiver operating up to 20 GHz. For receivers above this frequency the lowest noise system will again be achieved with a paramp and GaAs FET amplifiers can provide microwave IF amplifiers.

Improvements in transmitter EIRP, *eg* by the use of the multi-beam phased array transmitter, are continuing in parallel with developments in the low noise field. The aim is to reduce the overall cost of the communication system and for 076

systems that are intended to be used by many mobile users an increase in EIRP means a simpler and therefore lower cost receiver. Each system must be examined to give finally the most cost effective solution.

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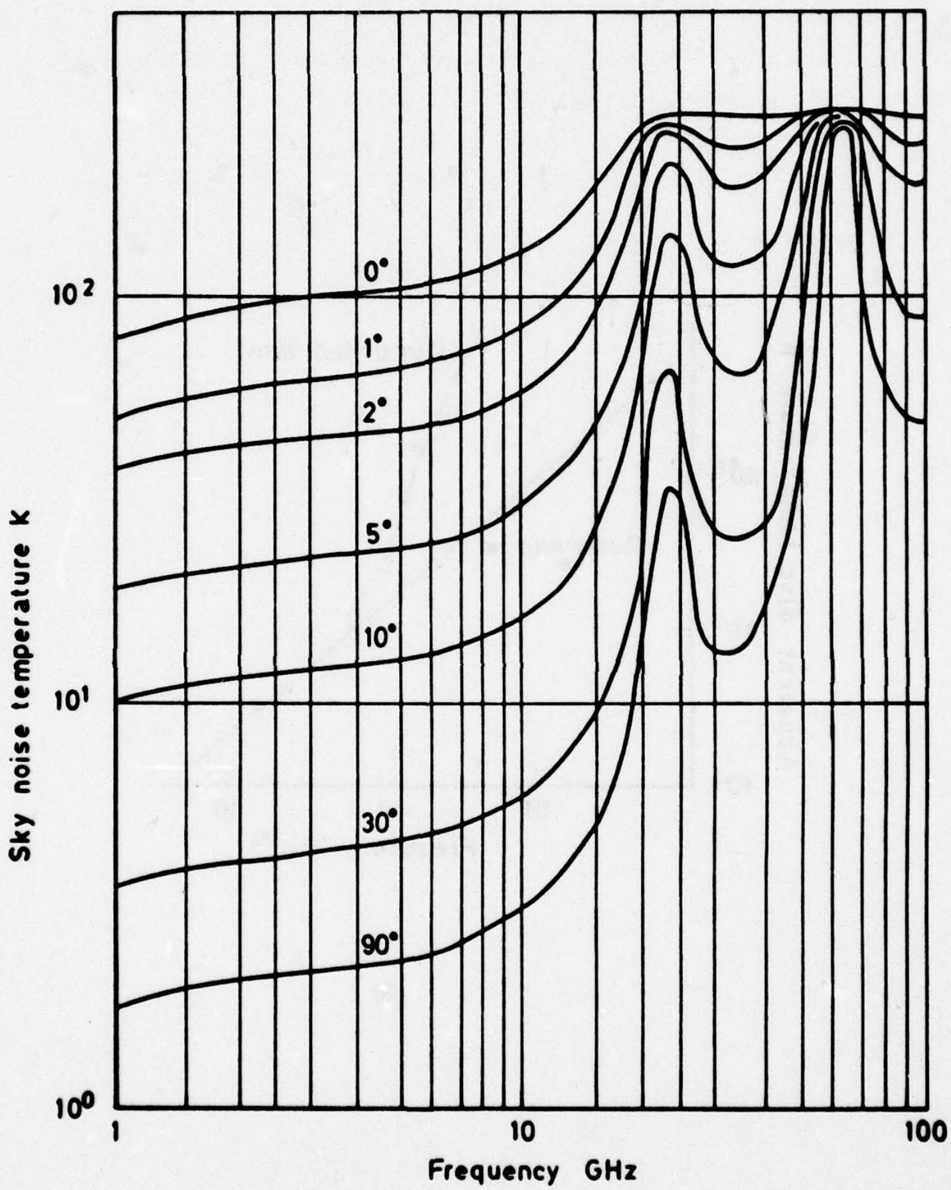

Fig 1 Sky noise as a function of frequency for different antenna elevation angles (after Hogg Ref 2)

TR 78076

Fig 2

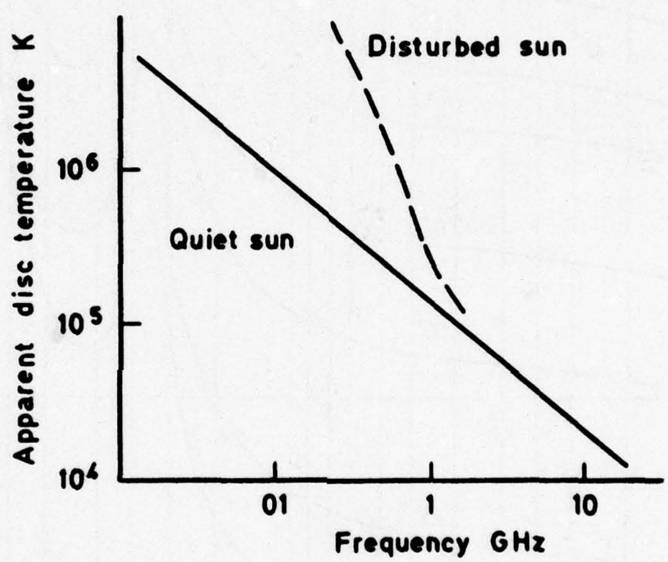

Fig 2 Apparent disc temperature of the sun as a function of frequency

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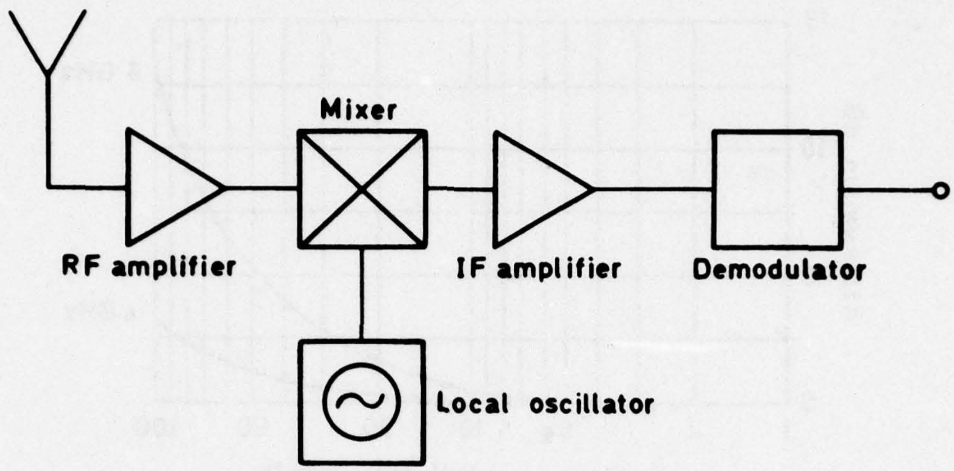


Fig 3 Heterodyne receiver

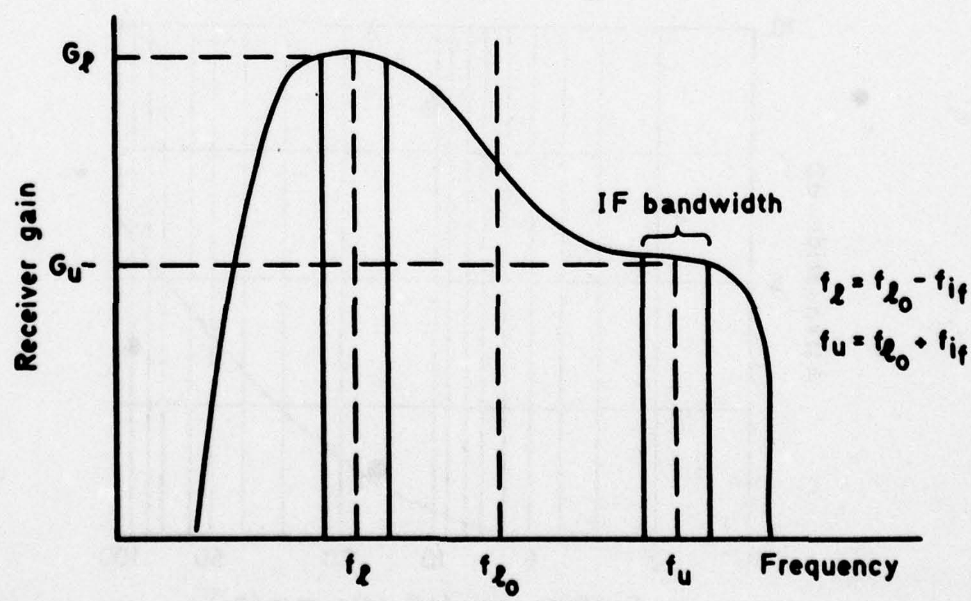


Fig 4 Receiver gain — frequency characteristic

Figs 5&6

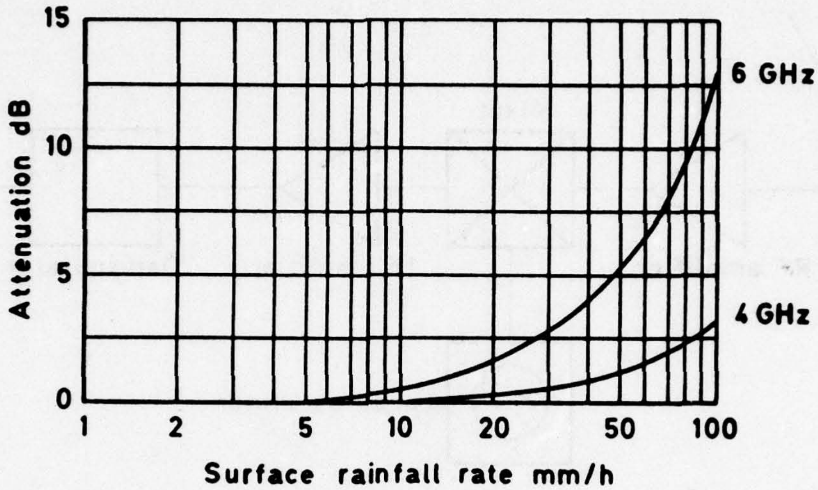


Fig 5 Attenuation due to absorption along a 5° elevated ray path as a function of surface rainfall rate

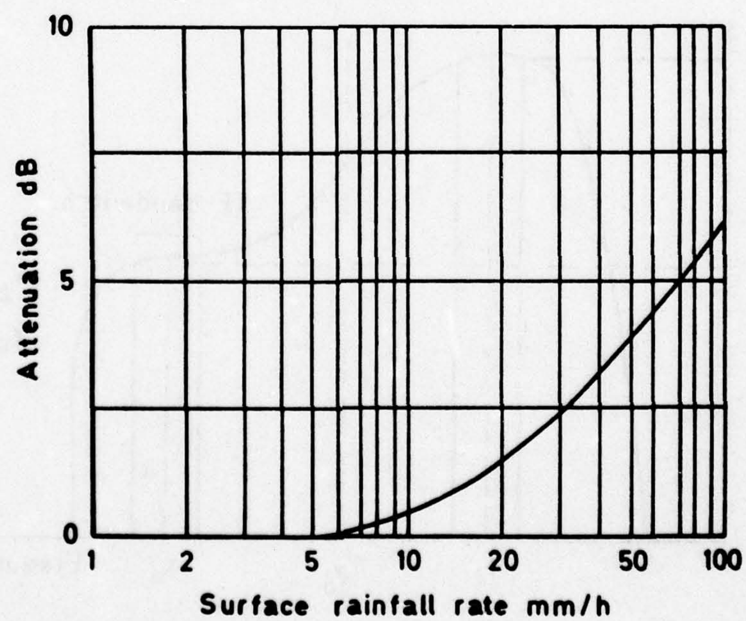


Fig 6 Relative noise temperature increase for a nominal 50 K noise temperature receiver in the absence of rain as a function of rainfall rate

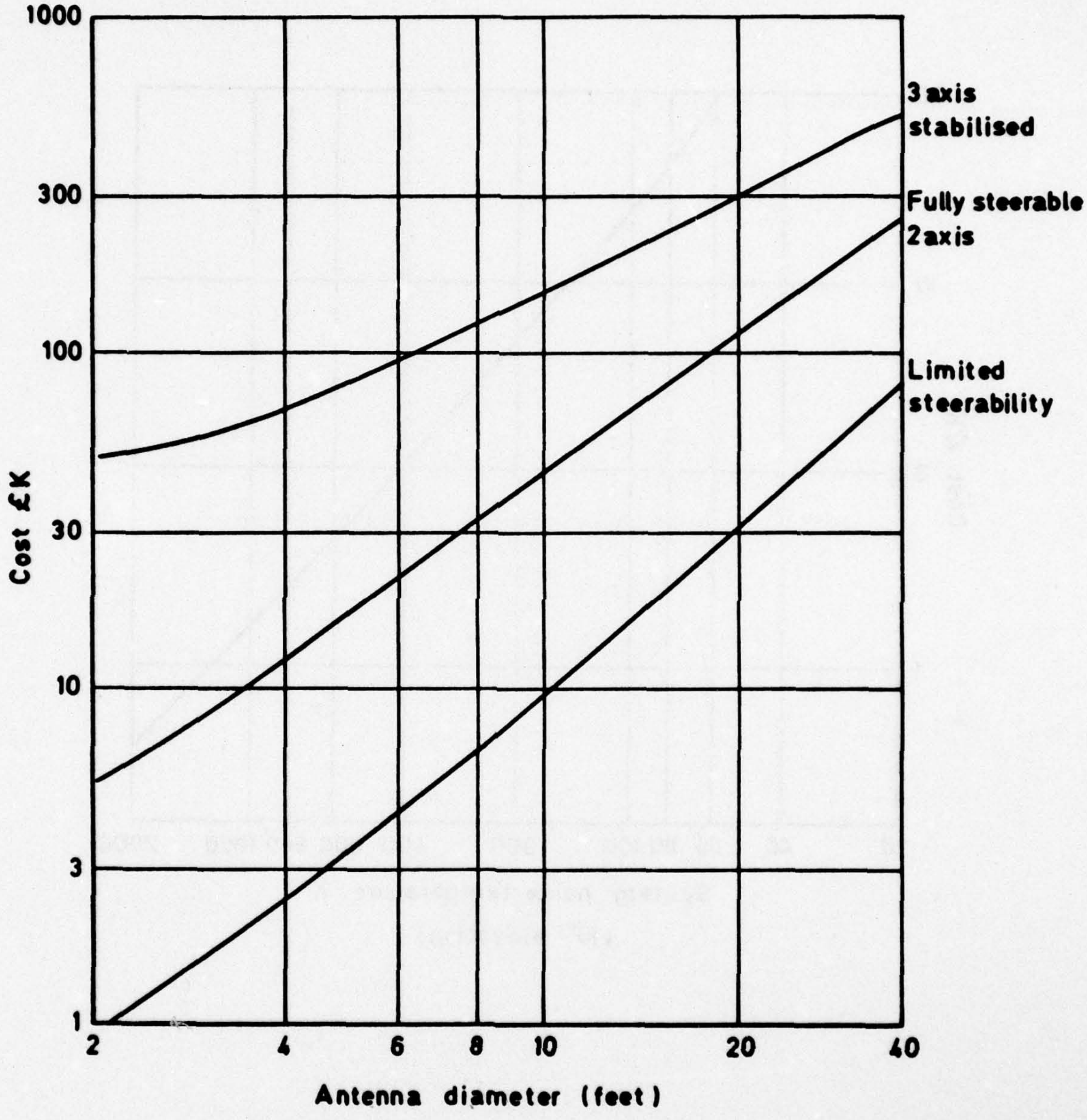


Fig 7 Antenna cost as a function of antenna diameter for three antenna mounting configurations

Fig 8

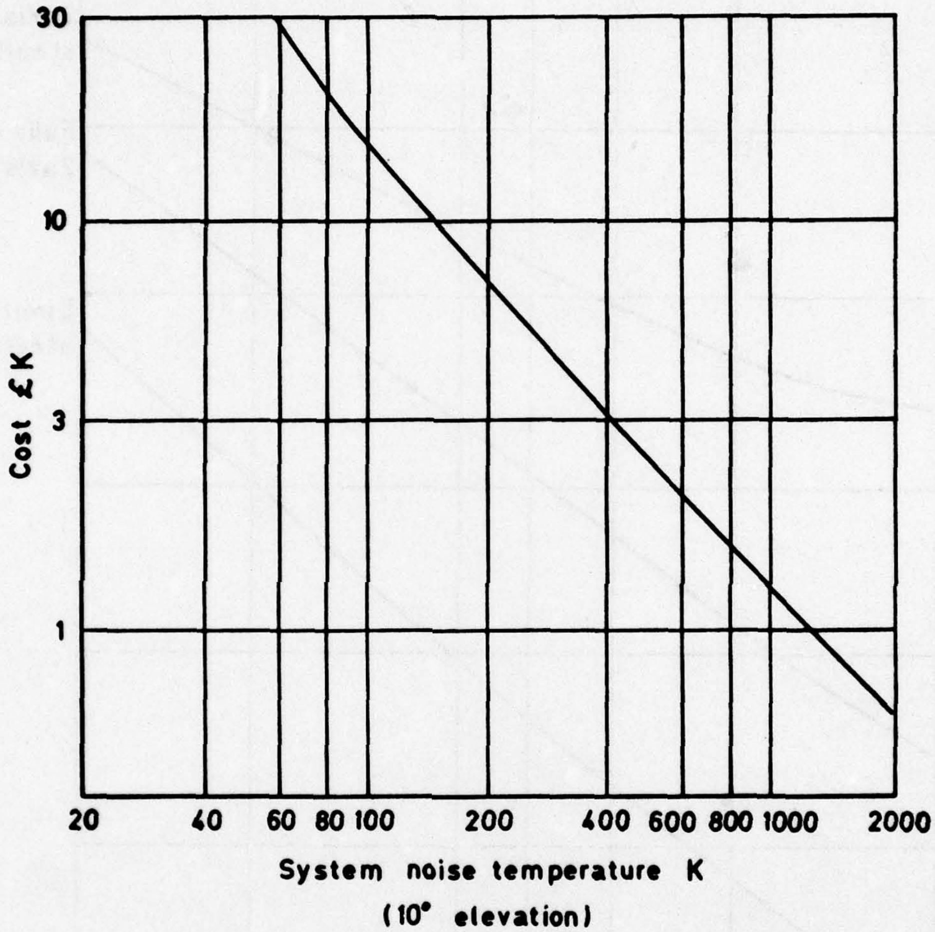


Fig 8 Approximate cost of RF amplifier units as a function of overall system noise temperature

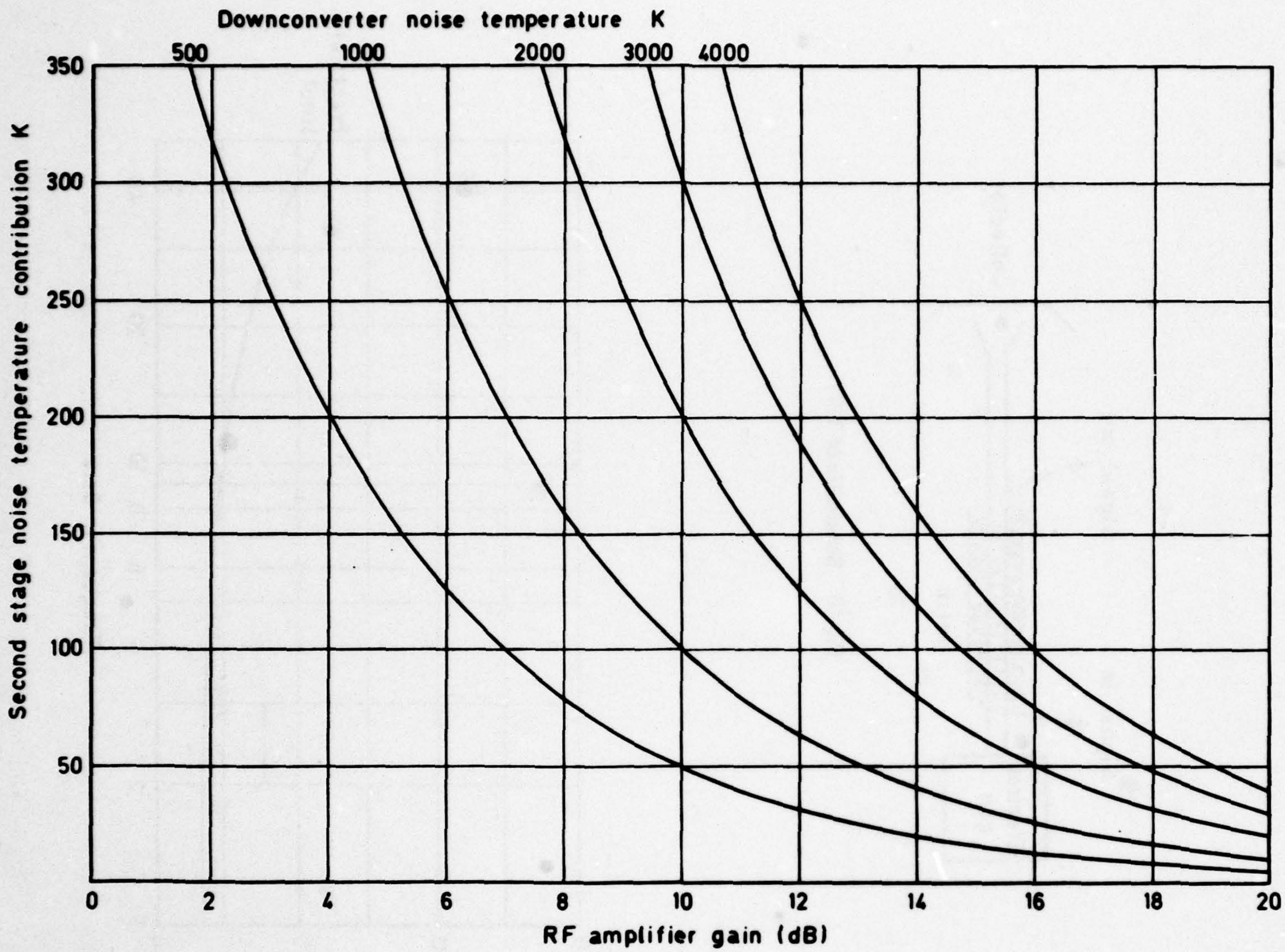


Fig 9 Downconverter noise temperature contribution to T_e as a function of RF amplifier gain for different values of downconverter noise temperature

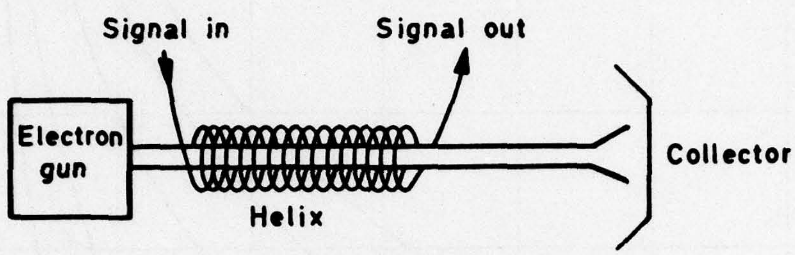


Fig 10 Schematic of TWT

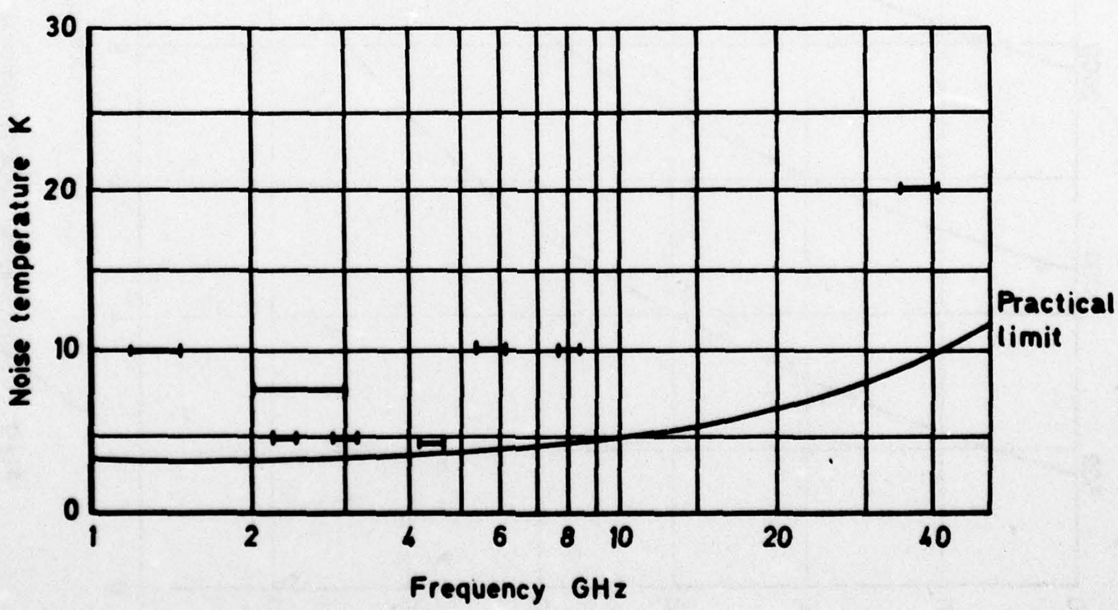
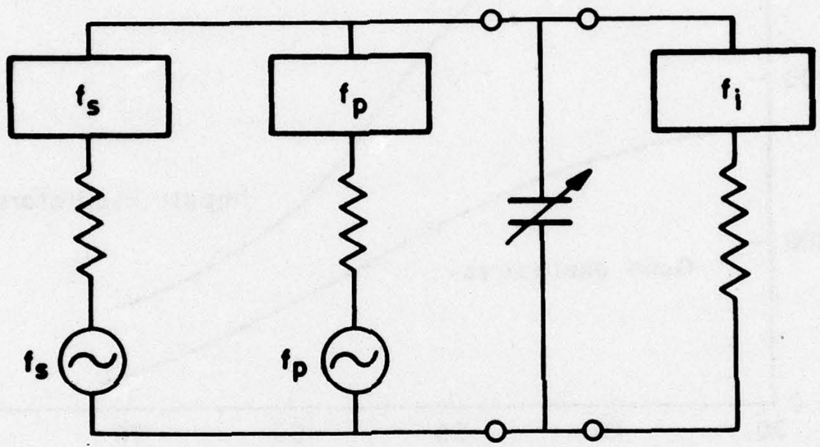


Fig 11 Noise temperatures of Masers as a function of frequency



f_s, f_p, f_i — filters tuned to signal, pump and idler frequencies respectively

Fig 12 Schematic of a three frequency parametric device

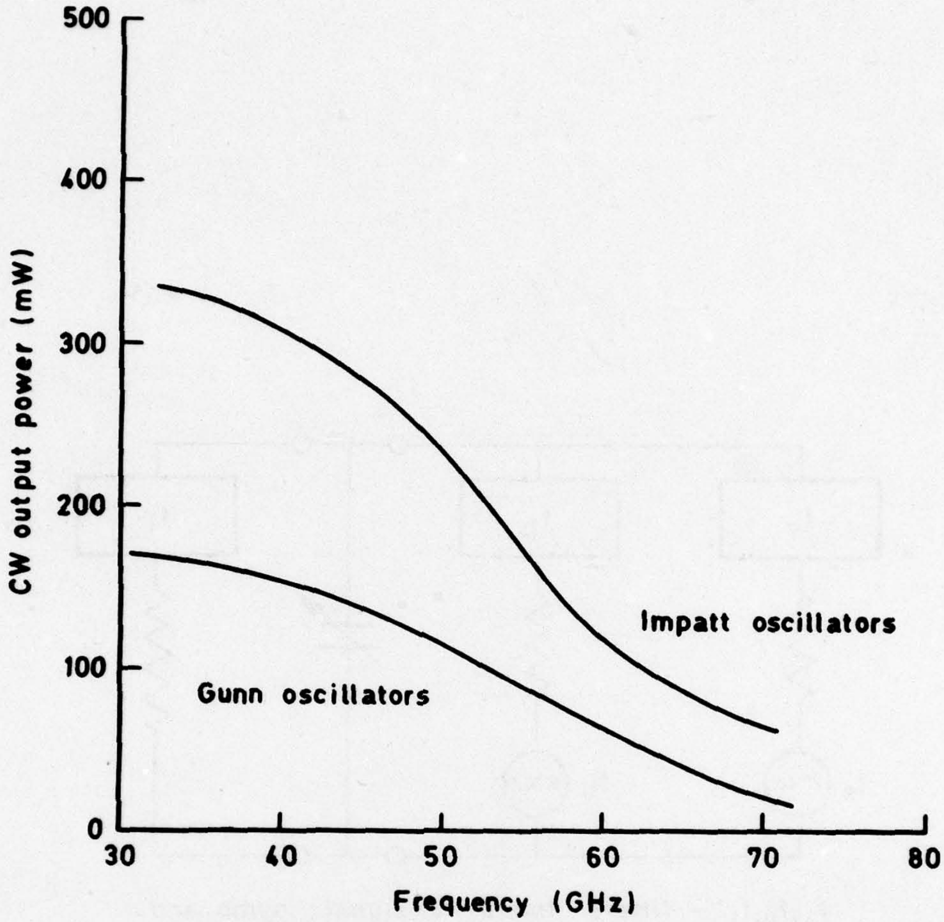


Fig 13 Available solid state oscillators

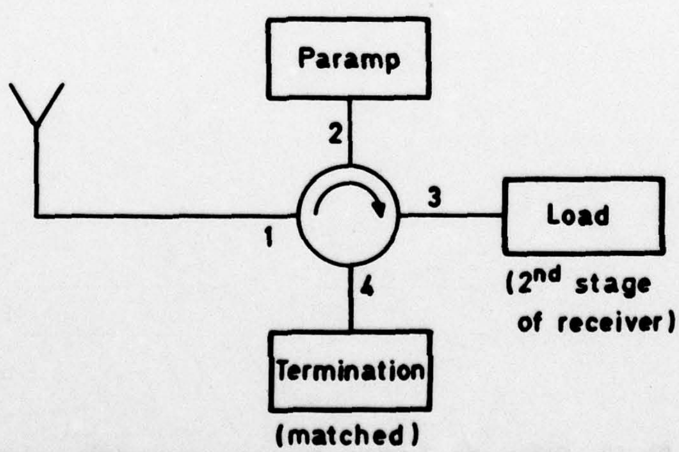


Fig 14 Circulator coupled Parametric Amplifier

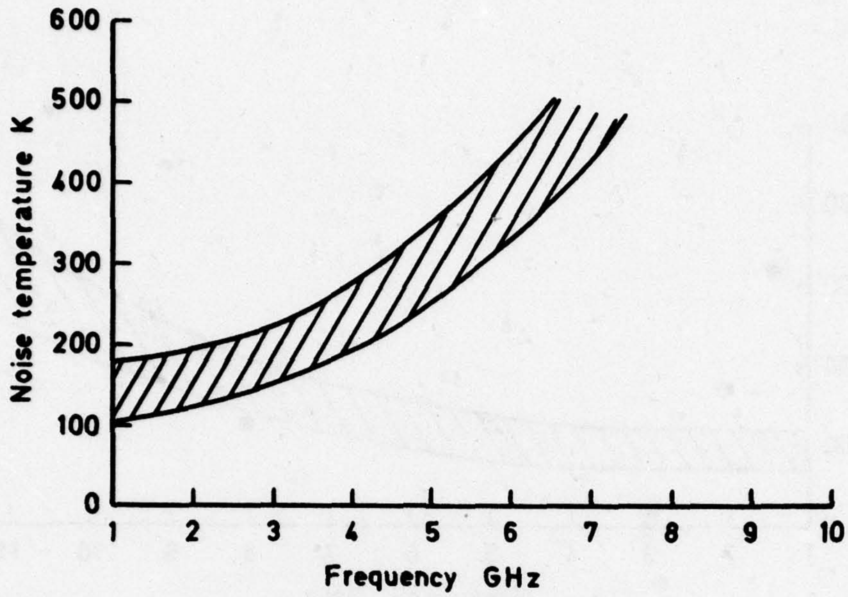


Fig 15 Noise temperature as a function of frequency for 1 μm emitter Si bipolar transistors

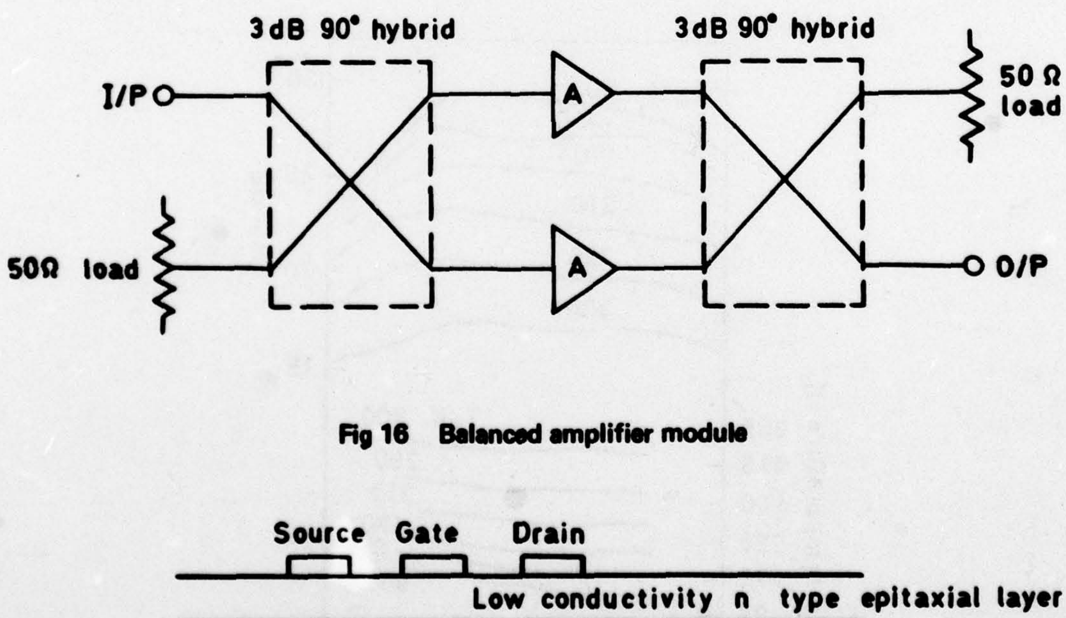


Fig 16 Balanced amplifier module

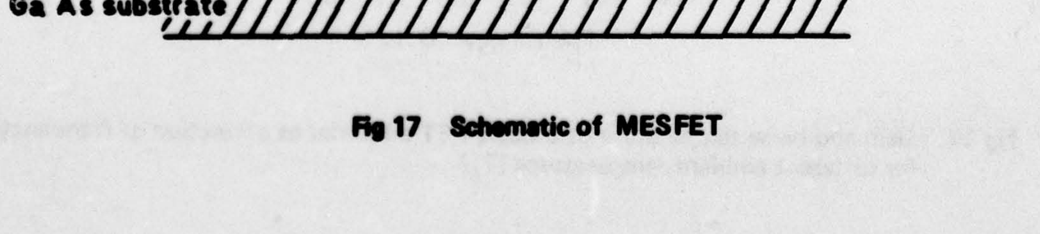


Fig 17 Schematic of MESFET

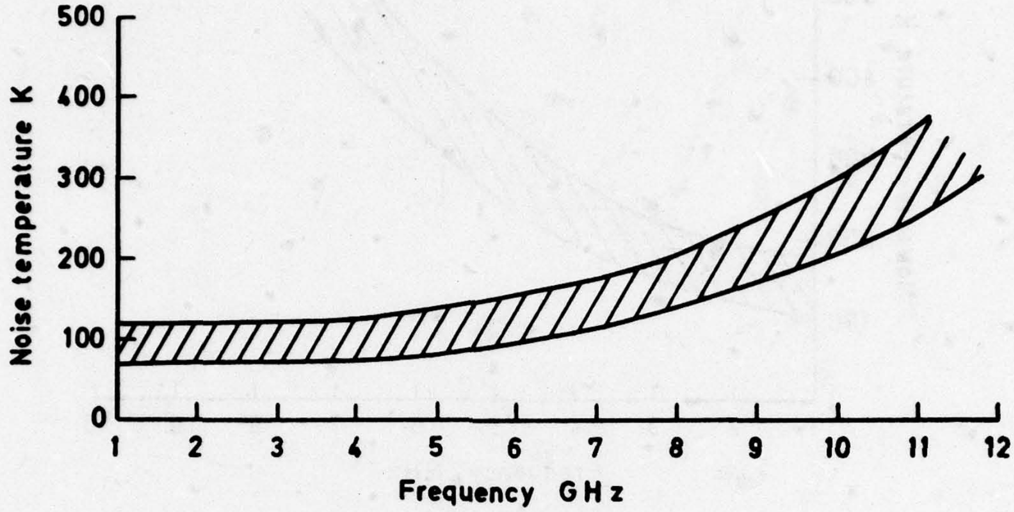


Fig 18 Noise temperature as a function of frequency for 0.5-1.0 μm gate length GaAs FETs

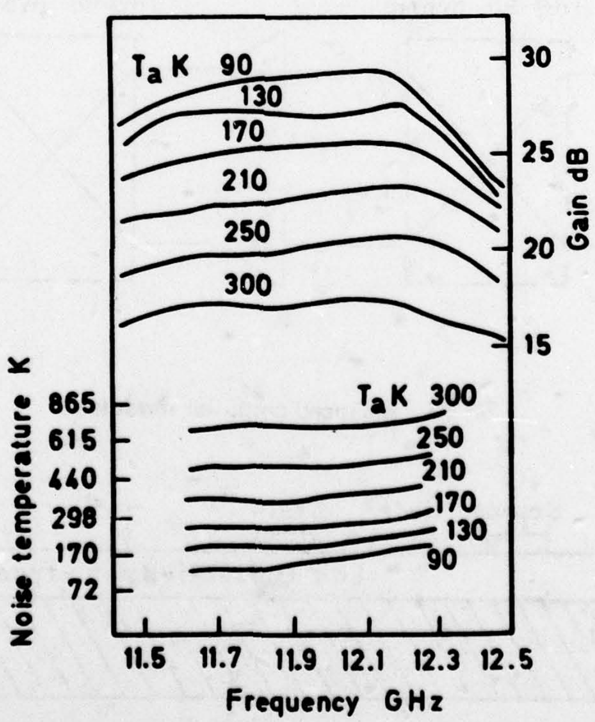
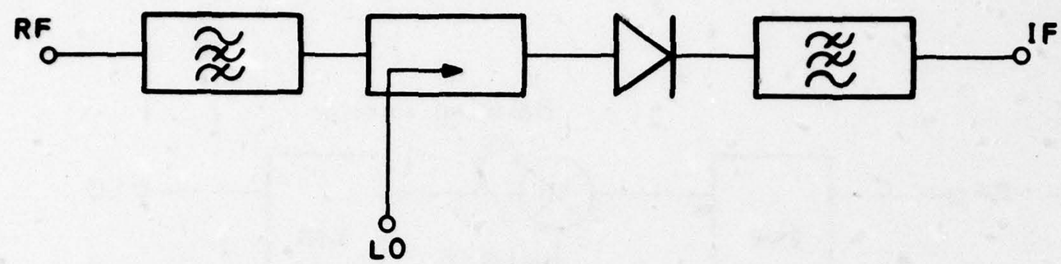
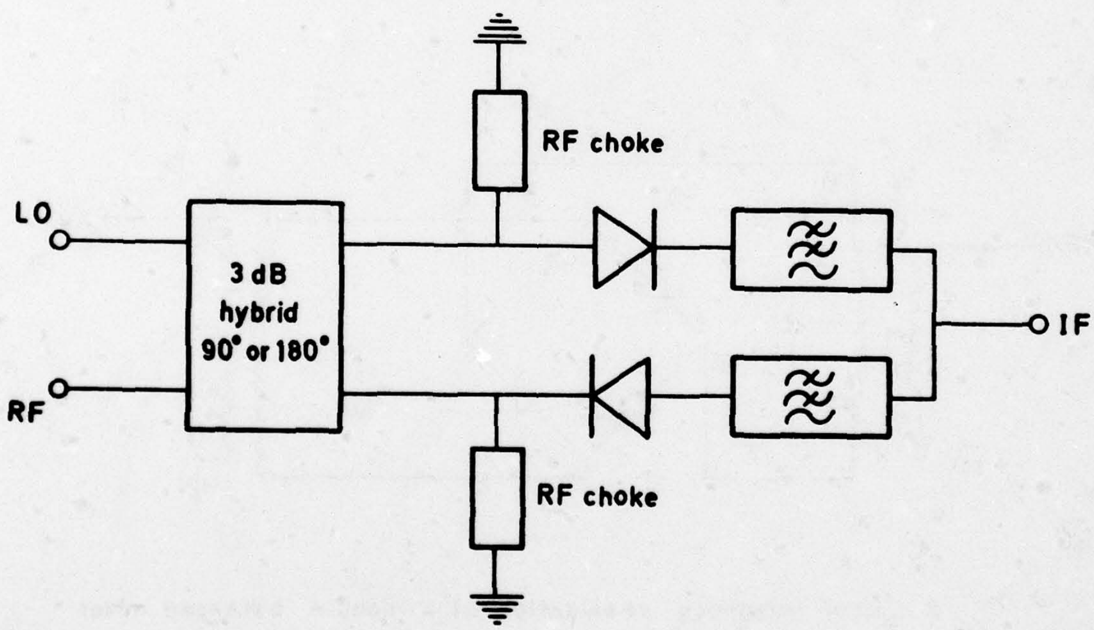


Fig 19 Gain and noise temperature of a GaAs FET amplifier as a function of frequency for different ambient temperatures (T_a)



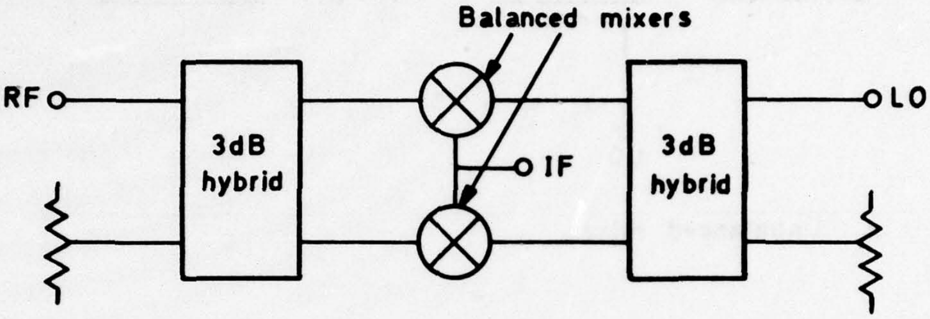
a Unbalanced mixer



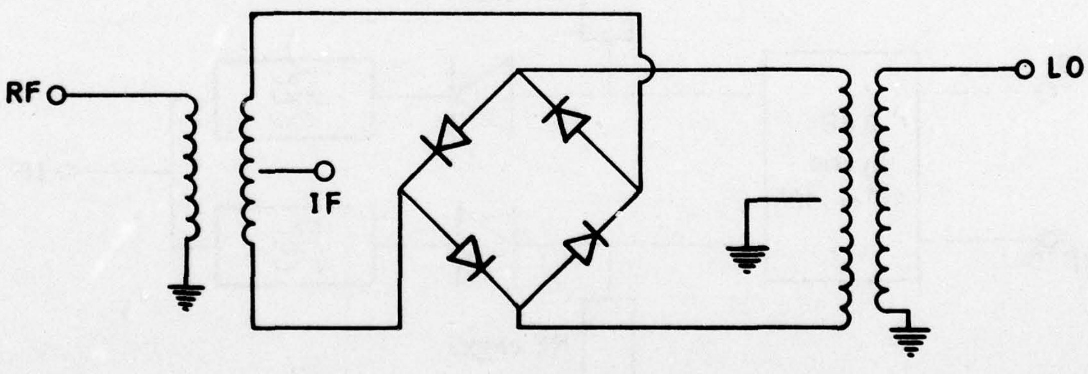
b Balanced mixer

Fig 20a&b Three basic Mixer types

Fig 20c&d



c Double balanced mixer



d Low frequency realisation of a double balanced mixer

Fig 20c&d Three basic Mixer types

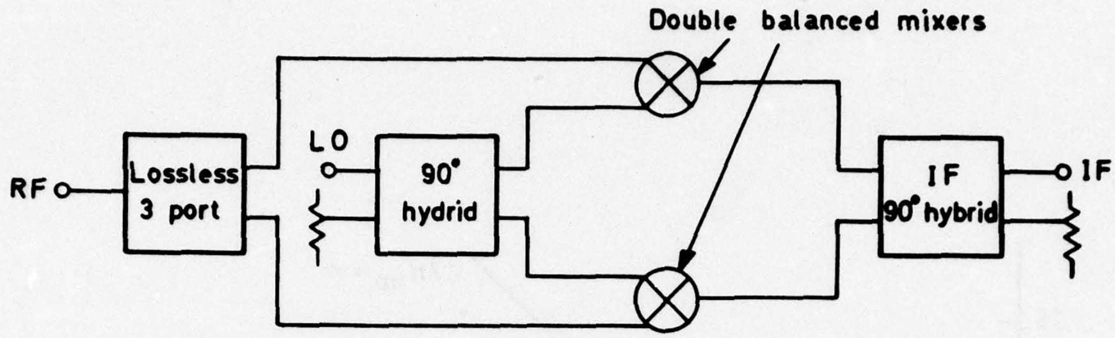


Fig 21 Image recovery Mixer

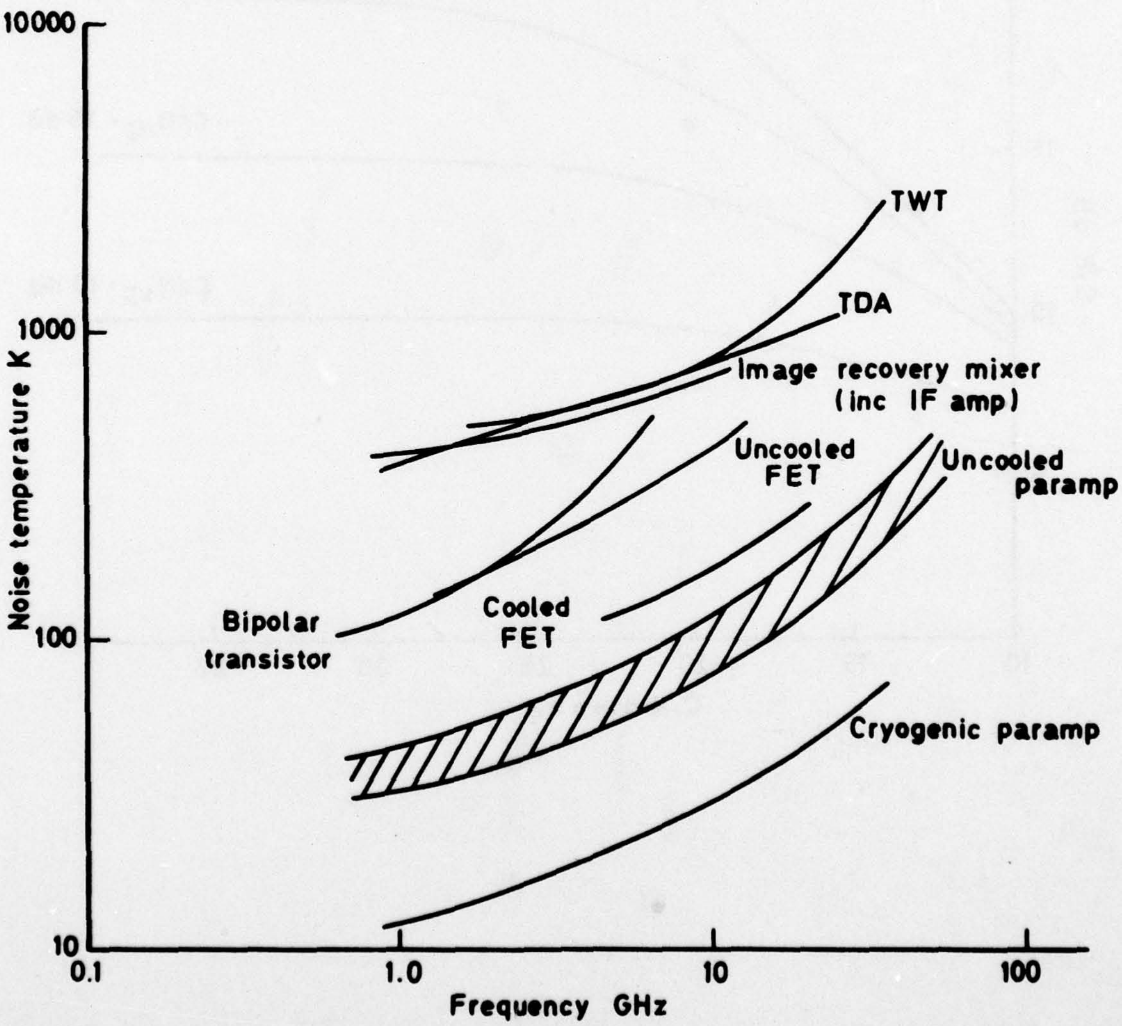


Fig 22 State of the art noise performance of front end devices as a function of frequency

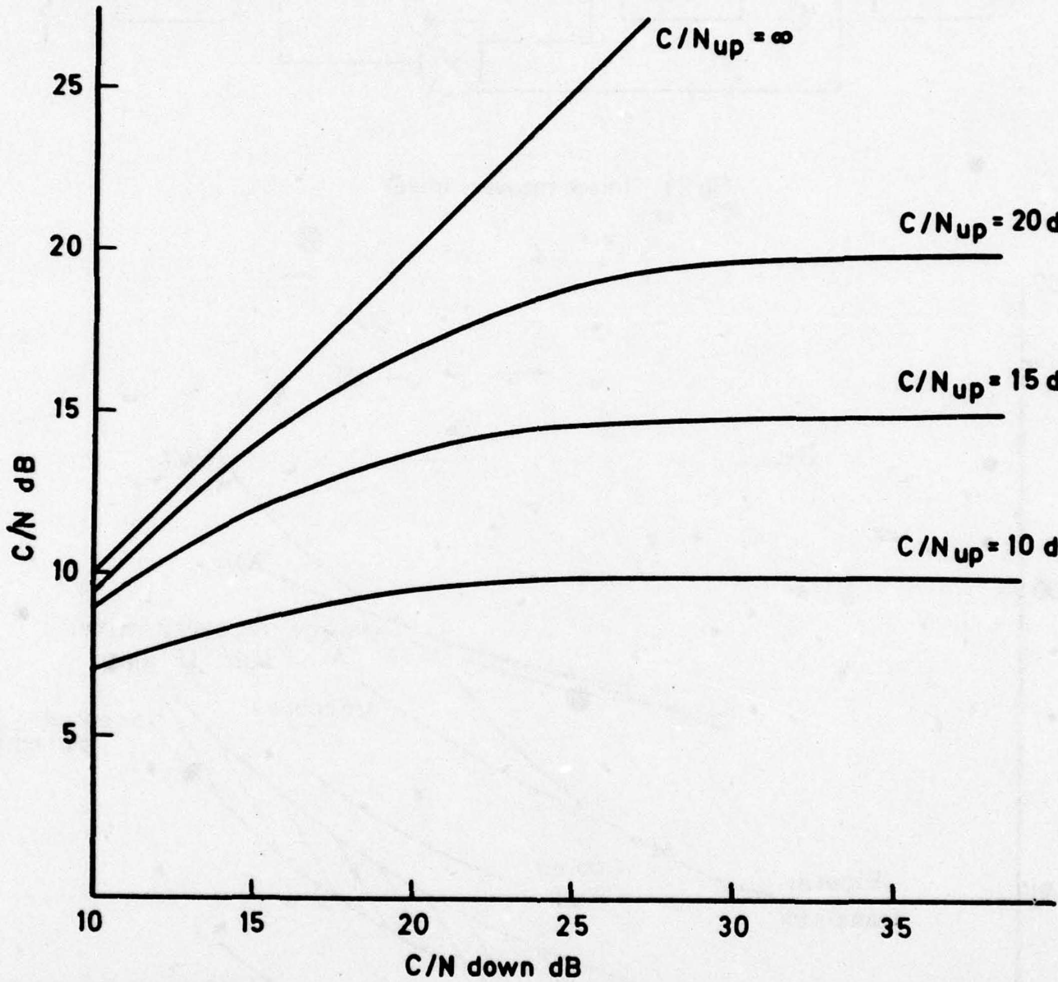


Fig 23 System C/N as a function of C/N_{down} for different values of C/N_u, the uplink and transponder C/N

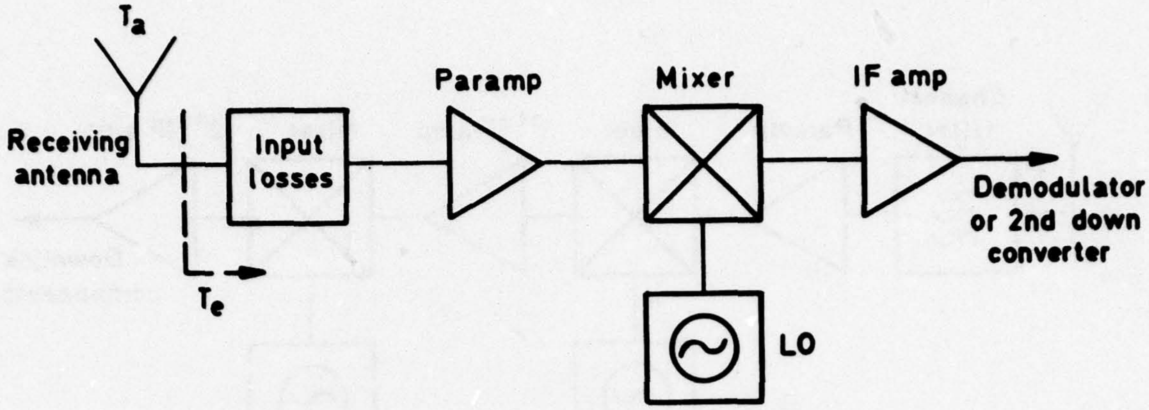


Fig 24 Typical 12GHz front end

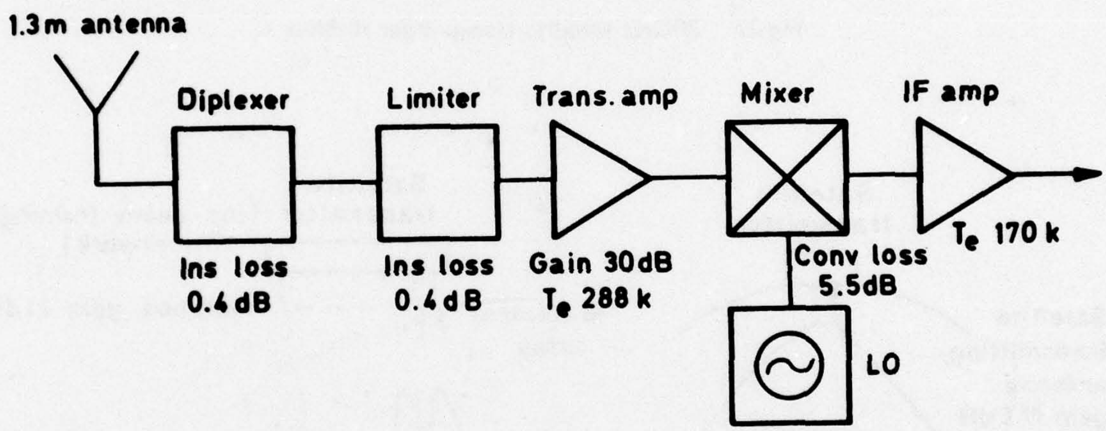


Fig 25 1.5GHz front end

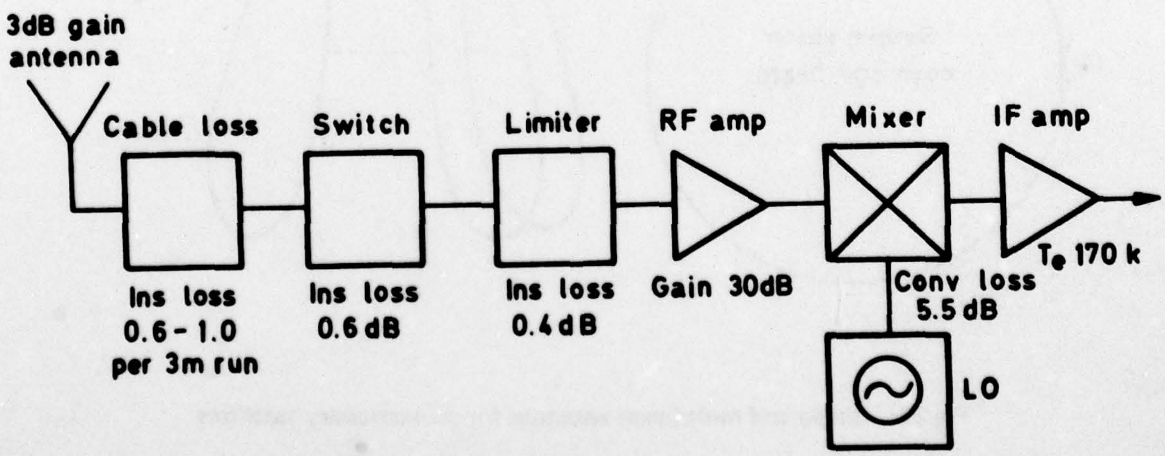


Fig 26 1.5GHz airborne receiver front end

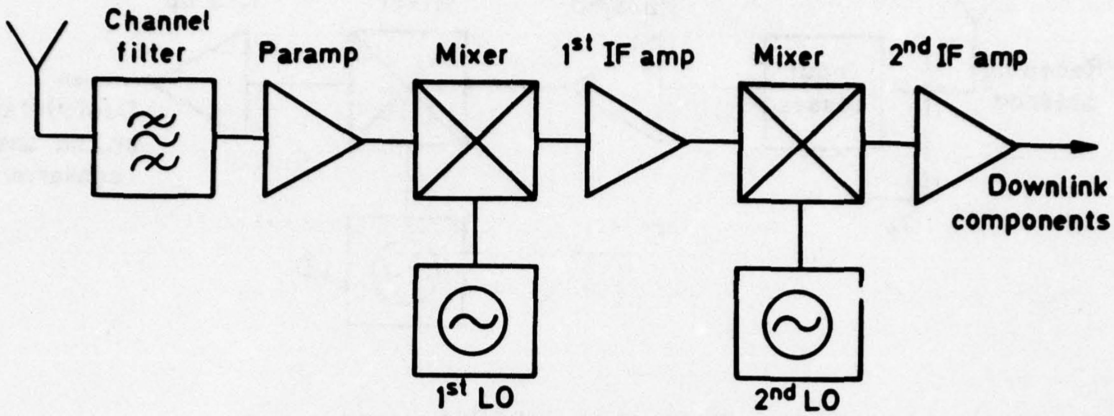


Fig 27 30GHz satellite transponder receiver

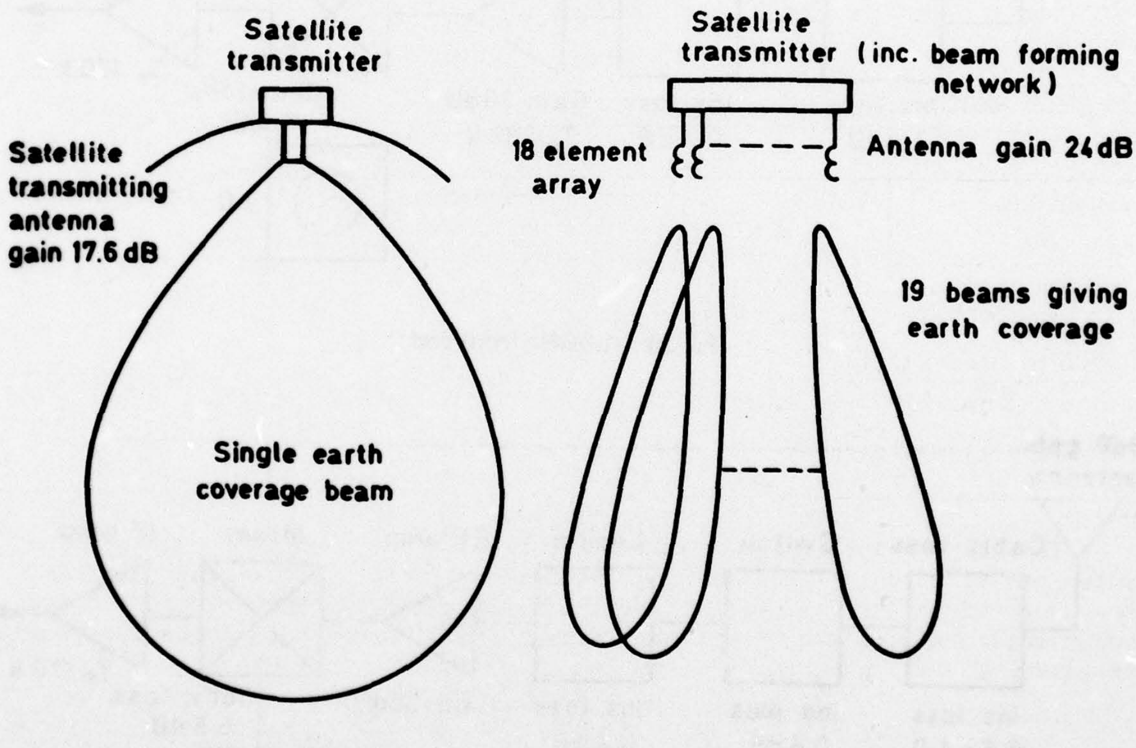


Fig 28 Single and multibeam antennas for geostationary satellites

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